SEPTEMBER · 1955

Proceedings OF THE IRE

Volume 43 Number 9

TESTING DEFLECTION YOKES

Syntronic Instruments, Inc.

stringent requirements of modern high-speed display systems demand careful king of deflection yoke characteristics. The yoke shown in the mirror above tube ing measured for transient response and resonant frequency.

IN THIS ISSUE

TABLE OF CONTENTS, INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 96A

IRE Standards on Industrial Electronics terms, Waveguide Definitions, and Receiver Testing appear in this issue

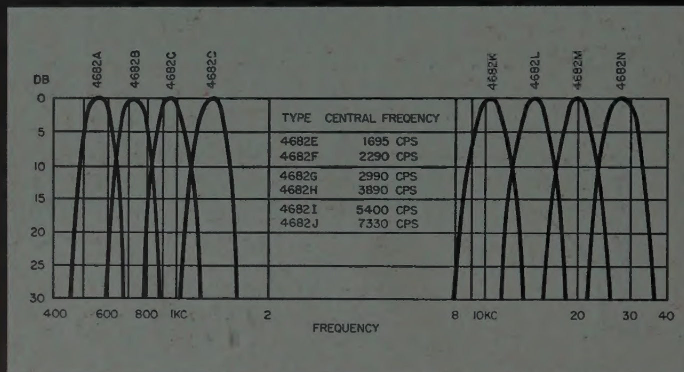
The Institute of Radio Engineers

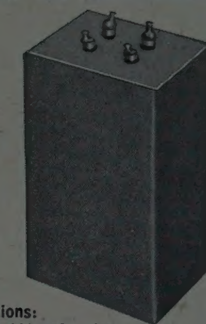
OUR MILLIONTH FILTER SHIPPED THIS YEAR...

FILTERS
FOR EVERY APPLICATION

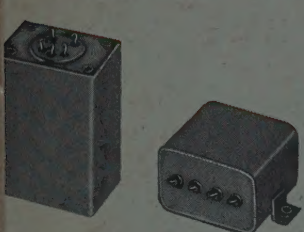
TELEMETERING FILTERS

UTC manufactures a wide variety of band pass filters for multi-channel telemetering. Illustrated are a group of filters supplied for 400 cycle to 40 KC service. Miniaturized units have been made for many applications. For example a group of 4 cubic inch units which provide 50 channels between 4 KC and 100 KC.

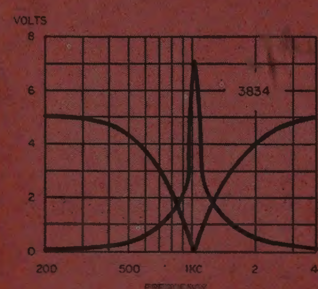


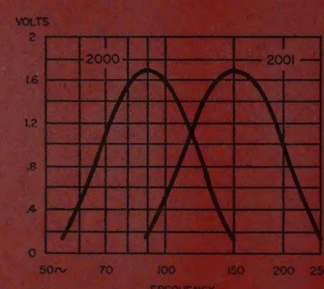


Dimensions:
(4682A) 1½ x 2 x 4″



Dimensions:
(3834) 1¼ x 1¾ x 2-3/16″.
(2000, 1) 1¼ x 1¾ x 1⅝″.





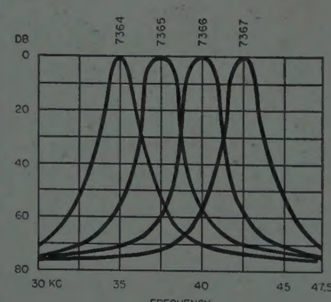
AIRCRAFT FILTERS

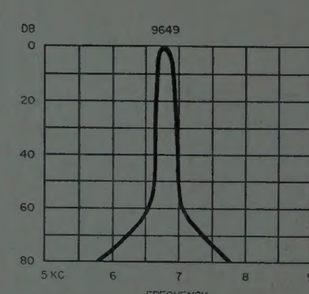
UTC has produced the bulk of filters used in aircraft equipment for over a decade. The curve at the left is that of a miniaturized (1020 cycles) range filter providing high attenuation between voice and range frequencies.

Curves at the right are that of our miniaturized 90 and 150 cycle filters for glide path systems.

CARRIER FILTERS

A wide variety of carrier filters are available for specific applications. This type of tone channel filter can be supplied in a varied range of band widths and attenuations. The curves shown are typical units.

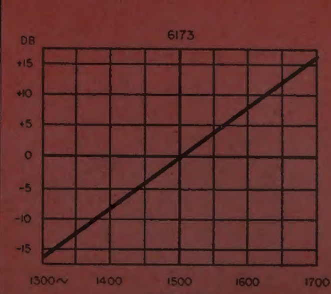


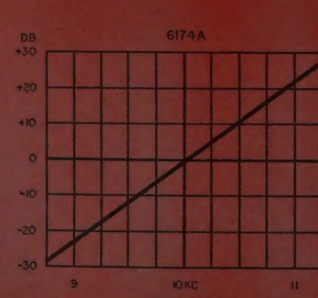


Dimensions:
(7364 series) 1⅝ x 1⅝ x 2¼″.
(9649) 1½ x 2 x 4″.

DISCRIMINATORS

These high Q discriminators provide exceptional amplification and linearity. Typical characteristics available are illustrated by the low and higher frequency curves shown.





Dimensions:
(6173) 1-1/16 x 1⅜ x 3″.
(6174A) 1 x 1¼ x 2¼″.

For full data on stock UTC transformers, reactors, filters, and high Q coils, write for Catalog A.

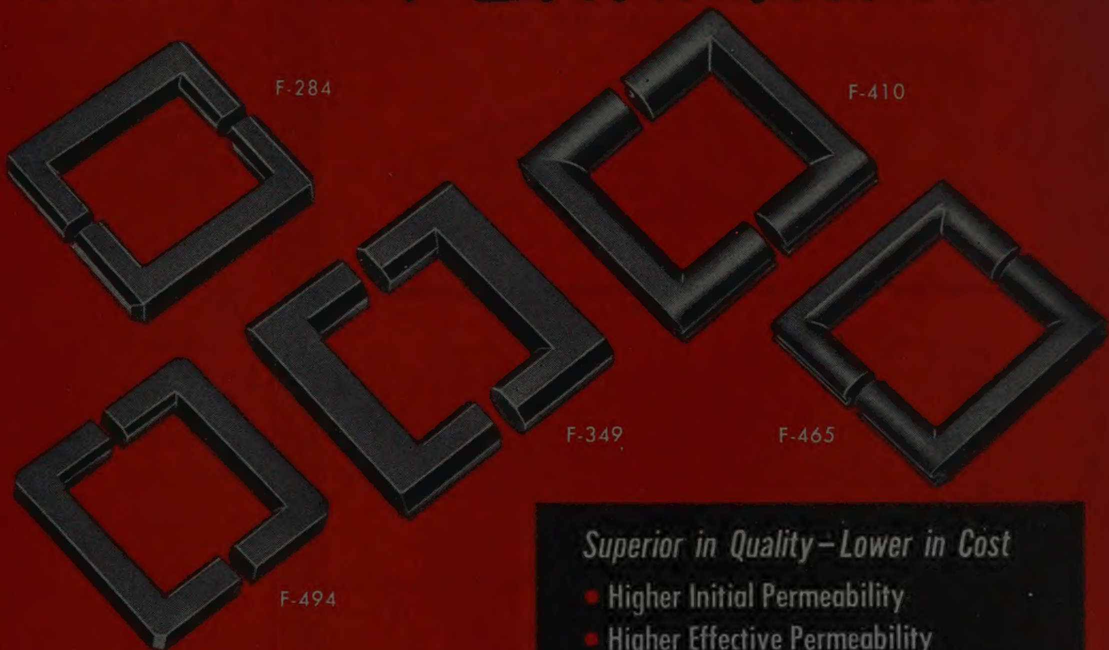
UNITED TRANSFORMER CO.

Another Advancement by

General Ceramics

Presenting—

SUPER FERRAMICS®



Superior in Quality—Lower in Cost

- Higher Initial Permeability
- Higher Effective Permeability at Higher Saturation Levels
- Lower Core Loss Resulting in Less Temperature Rise
- Greater Uniformity Through Improved Production Techniques

New Super-Grade Ferrites from the Laboratories of General Ceramics—

New Super-Ferramics are magnetic ferrites with properties once considered beyond the realm of achievement. The first of this series Ferramic O, (see property chart) has been released and is now available in production quantities. Engineers and product designers are invited to request complete information on Ferramic O. Call or write for data today!

MAGNETIC PROPERTIES OF FERRAMIC O-1

PROPERTIES	UNIT	FERRAMIC O-1
Mu_o at 50 kcs.	—	1200
Mu_{max}	—	6000
Saturation Flux Density B_s	Gauss	4100
Residual Magnetism B_r	Gauss	2500
Coercive Force H_c	Oersteds	0.20
Curie Temperature	+ °C.	165
Volume Resistivity	—	Low
Loss Factor at 50 kcs.	$\frac{1}{u_0 Q}$	0.000010
Temp. Coeff. of Initial Perm. (50 Kcs)	%/°C.	+0.75

GENERAL CERAMICS CORPORATION STEATITE INSULATORS

General CERAMICS CORPORATION

TELEPHONE: VALLEY 6-5100

GENERAL OFFICES and PLANT: KEASBEY, NEW JERSEY

MAKERS OF STEATITE, ALUMINA, ZIRCON, PORCELAIN, SOLDERSEAL TERMINALS, "ADVAC" HIGH TEMPERATURE SEALS, CHEMICAL STONEWARE, IMPERVIOUS GRAPHITE, FERRAMIC MAGNETIC CORES

Meetings with Exhibits

● **As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups which include exhibits.**

Δ

Sept. 12-16, 1955
Tenth Annual Instrument Conference & Exhibit, Shrine Exposition Hall & Auditorium, Los Angeles, Calif.
Exhibits: Mr. Fred J. Tabery, 3442 So. Hill St., Los Angeles 7, Calif.

Sept. 26-27, 1955
IRE Sixth Annual Meeting of the Professional Group on Vehicular Communications, Hotel Multnomah, Portland, Ore.
Exhibits: Mr. Henry S. Broughall, General Electric Co., 2727 N.W. 29th Ave., Portland, Ore.

October 3-5, 1955
National Electronics Conference, Sherman Hotel, Chicago, Ill.
Exhibits: Mr. G. J. Argall, c/o DeVry Technical Institute, 4141 Belmont Ave., Chicago 41, Ill.

Oct. 31-Nov. 1, 1955
IRE East Coast Conference on Aeronautical & Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md.
Exhibits: Mr. C. E. McClellan, Westinghouse Electric Corp., Air Arm Div., Friendship International Airport, Baltimore, Md.

Nov. 3-4, 1955
Annual Electronics Conference, Kansas City Section, Town House Hotel, Kansas City, Kans.
Exhibits: Mr. Charles V. Miller, Bendix Aviation Corp., P.O. Box 1159, Kansas City 41, Mo.

Nov. 7-9, 1955
Eastern Joint Computer Conference (IRE-AIEE-ACM), Hotel Statler, Boston, Mass.
Exhibits: Mr. J. D. Porter, Digital Computer Lab., Barta Building, M.I.T., Cambridge, Mass.

Nov. 28-30, 1955
Instrumentation Conference & Exhibit, Atlanta Biltmore Hotel, Atlanta, Ga.
Exhibits: Mr. W. B. Wrigley, Engineering Experiment Station, Georgia Institute of Technology, Atlanta, Ga.

December 12-16, 1955
EJC Nuclear Science and Engineering Congress, Cleveland, Ohio
Exhibits: Engineers Joint Council, 33 W. 39th St., New York, N.Y.

Feb. 9-11, 1956
Eighth Annual Southwestern IRE, Conference and Electronics Show, Municipal Auditorium, Oklahoma City, Okla.
Exhibits: Mrs. Charles E. Harp, P.O. Box 764, Oklahoma City, Okla.

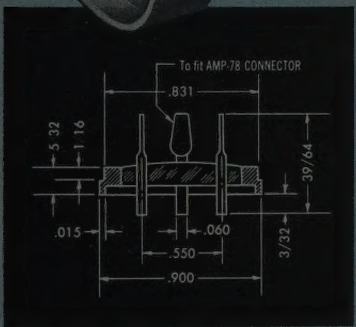
March 19-22, 1956
IRE National Convention and Radio Engineering Show, Waldorf-Astoria Hotel and Kingsbridge Armory and Palace, New York, N.Y.
Exhibits: Mr. William C. Copp, Institute of Radio Engineers, 1475 Broadway, New York 36, N.Y.

Now! For AMP Taper Pin Connectors

E-I COMPRESSION HEADERS

Type 90 GS/60W-AMP/S Compression Header, available with from 8 to 14 terminals, shown four times actual size.

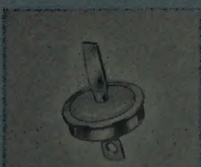
...offering fast connect and disconnect, speedy assembly and positive connections without soldering!



Available

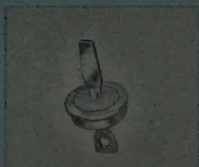
IN COMPRESSION HEADERS AND PRACTICALLY ALL STANDARD E-I SINGLE TERMINAL EYELETS—

E-I offers single and multiple terminal type compression headers and practically every standard E-I single terminal compression eyelet for use with Type 78 AMP connectors*. For recommendations on specific sealed terminal applications, consult an E-I sales engineer, today!

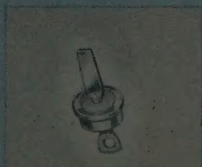


DAS/60W-AMP/S COMPRESSION TYPE (Actual Size)

FAS/60W-AMP/S COMPRESSION TYPE (Actual Size)



CCAS/60W-AMP/S COMPRESSION TYPE (Actual Size)



ABAS/60W-AMP/S COMPRESSION TYPE (Actual Size)

(Actual Size)

90GS/60W-AMP/S-14A COMPRESSION TYPE

*Products of Aircraft-Marine Products, Inc. of Harrisburg, Pa.

Division of Amperex Electronic Corporation

EI ®

ELECTRICAL INDUSTRIES

44 SUMMER AVENUE, NEWARK 4, NEW JERSEY

NEWS and NEW PRODUCTS

September 1955

Standard Ratio Transformer

Four new ruggedized standard ratio transformers have been added to the line of precision ac voltage dividers developed by **Gertsch Products, Inc.,** 11846 Mississippi Ave., Los Angeles 25, Calif.

The PT Series consists of nine models of both rack mounted and case models, specifically designed to divide ac voltage with accuracies as good as 0.005 per cent and resolution as good as 0.00001 per cent. Models are available to cover frequencies from 30 to 3,000 cps (to 10,000 cps at reduced accuracy).

The four new models have ruggedized heavy silver rotary switches, for use wherever severe continuous service is required. According to the manufacturer the permanent "built-in" accuracy contained in all models, is not subject to the variations normally experienced with resistive dividers.

Uses for the standard ratio transformers include core material investigation, ac meter calibration, transformer turns investigation, checking resolvers, servos, computers, synchros, selsyns, and ac transducers, bridge ratio arm, ac potentiometer checking, and as a ratio standard.

LF Q Meter

The Kilo-Q, most recent addition to its line of electronic instruments was announced recently by **Kay Electric Co.,** 14 Maple Ave., Pine Brook, N. J.

A low frequency Q meter, the Kilo-Q will cover a range of 20

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

cps to 1 mc. For ease of operation, the unit combines a direct reading dial over the entire range, accurate to 1 per cent.

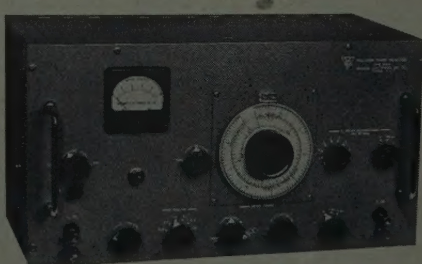
Two positions are provided on the Q Range Control with full scale Q readings of 0 to 125 and 0 to 250, respectively. For reading Q values between 250 and 500, it is necessary to set the lever control to the X2 position on the dial and double the reading on the 250 scale.

The instrument provides a tuning capacitor range of 60–1,200 $\mu\mu$f. A vernier capacitor control in shunt is provided to facilitate tuning of sharp "Qs."

Oscillator frequency accuracy, 2 to 5 per cent. Calibration capacitance accuracy, 1 per cent. Price $695 f.o.b. plant. For complete information, write the manufacturer.

Precision Phase Detector

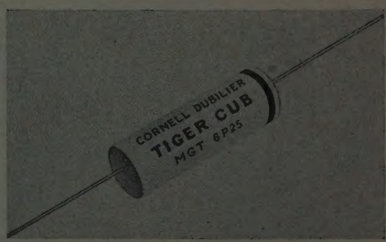
This instrument manufactured by **Advance Electronics Co., Inc.,** 451 Highland Ave., Passaic, N. J., will measure time delay, phase delay, or envelope delay with error less than 1 per cent or 0.1° between two alternating voltages from 10 kc up to 15 mc. Essentially Type 205a precision phase detector consists of two input amplifiers, a continuously variable delay line, a step variable delay line, a differential tuned amplifier, a balanced phase detector, and a sensitive output indicator.



The smallest time delay that can be read on the dial is 5×10^{-10} seconds; the smallest phase angle in degrees that can be read on the dial is equal to $5 \times 10^{-10} \times 36 \times$ frequency in cps. The frequency range is 10 kc to 15 mc. The time delay of the step variable delay line is 5 μs in step of 0.05 μs. Three plug-in units of continuously variable delay lines are supplied with the instrument, 0 to 0.4 μs, 0 to 0.25 μs, and 0 to 0.05 μs. The maximum phase range that can be measured with the instrument is equal to the total time delay of the continuously variable delay line plus the step variable delay line multiplied by the frequency of the signals and 360. The indicator sensitivity is approximately 0.01 volt full scale maximum without the probe, and 0.1 volt with the probe. Two low capacity probes with input capacitance less than 4 $\mu\mu$f are supplied with the unit. The panel binding post has about 1 megohm shunted with 12 $\mu\mu$f on both input channels.

Paper Dielectric Tubulars

Cornell-Dubilier Electric Corp., South Plainfield, N. J., announces the development of its new "Tiger Cub" type MGT, high temperature paper dielectric tubular capacitors. This new capacitor is designed to operate effectively at temperatures from −55°C to +100°C.



The capacitance stability of the new "Tiger Cub" is such that it varies less than 10 per cent over this temperature range. Longer service life is assured by Vikane impregnation. An external wax dip provides added moisture protection that will withstand 250 hours of continuous exposure in 90 per cent relative humidity at 40°C.

The "Tiger Cub" MGT paper tubular capacitors are available in capacities from 0.001 $\mu\mu$f to 1.0 $\mu\mu$f in 6 voltage ranges from 100 to 1,600 volts dc working. Low resistance lead wires are soldered to extended foils and held firmly in place by Polykane, the high temperature, non-melting end fill. Request Bulletin 168.

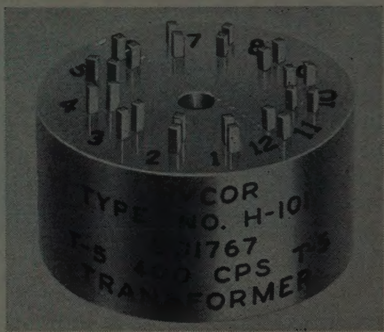
(Continued on page 16A)

News-New Products

(Continued from page 14A)

Minature Power Transformers

Hycor Company, Inc., 11423 Vanowen St., North Hollywood, Calif., announces a new line of miniature power transformers for 400 cps and higher frequencies. The units are available with output power ratings up to 15 va with



multiple windings from 1 volt to 500 volts. They are available in miniature metal cases or in plastic encapsulated form to satisfy MIL-T-27 requirements. The torodoil construction minimizes external fields and results in extremely high efficiency. Bulletin WT lists stock types and is available upon request.

Terminal Catalog

Hermetic Seal Products Co., 29 South Sixth St., Newark 4, N. J. announces the availability of their new 4-page brochure on Vac-Tite Compression, Single Terminal Feed-Thru's and Stand-Offs. This new 4-page brochure provides industry with a coordinated standardization of single terminal feed-thru's and stand-off's. The parts illustrated in the new brochure are of Vac-Tite construction, an exclusively developed glass-to-metal chemically bonded compression construction.

The brochure introduces a wide variety of specially designed flanged bodies offered to industry for use in projection weld assembly, for soldering to curved surfaces, and other special applications. In addition, designs capable of withstanding extremely high pressures are available in flanged or threaded bodies. Single terminal types have been developed by Hermetic that incorporate space within the seal for mounting small components. This type requires a flat plate for closure.

(Continued on page 181A)

Redesigned

to solve your horizontal deflection problems

CIRCUIT DESIGNED

OK

CIRCUIT TESTED

—these Sylvania deflection amplifier tubes offer higher plate currents, greater dissipation

Here is a full line of Sylvania Tubes—made to take the tighter conditions of horizontal deflection circuits in streamlined TV chassis designs.

New plate and grid designs achieve minimum zero bias plate to screen grid current ratios of 10 to 1. Plate dissipation has been increased to provide more stable performance throughout tube life. Designed to exhibit low plate knee characteristics, these tubes eliminate "snivet" problems when operated properly within ratings.

Whatever the nature of your TV design problem, Sylvania Tubes are "circuit-designed and circuit-tested" to meet your needs.

Deflection Types for Transformer Circuits

6BQ6GTA 6CD6GA
6CU6 6DQ6

Deflection Types for Series—String circuits

12BQ6GTA 12CU6
25BQ6GTA 25CD6GA
12DQ6

SYLVANIA®

SYLVANIA ELECTRIC PRODUCTS INC.
1740 Broadway, New York 19, N. Y.
In Canada: Sylvania Electric (Canada) Ltd.
University Tower Building, Montreal

SYLVANIA ELECTRIC PRODUCTS INC.
Dept. I32P, 1740 Broadway, New York 19, N. Y.

☐ Please send complete data on "circuit-designed and circuit-tested" deflection amplifier types.

Check other tube interests.

☐ Other entertainment types ☐ Control equipment types
☐ Military types ☐ Test equipment types
☐ Special-Purpose types ☐ ________types

Name________________

Address________________

City________________Zone____State________

LIGHTING • RADIO • ELECTRONICS • TELEVISION • ATOMIC ENERGY

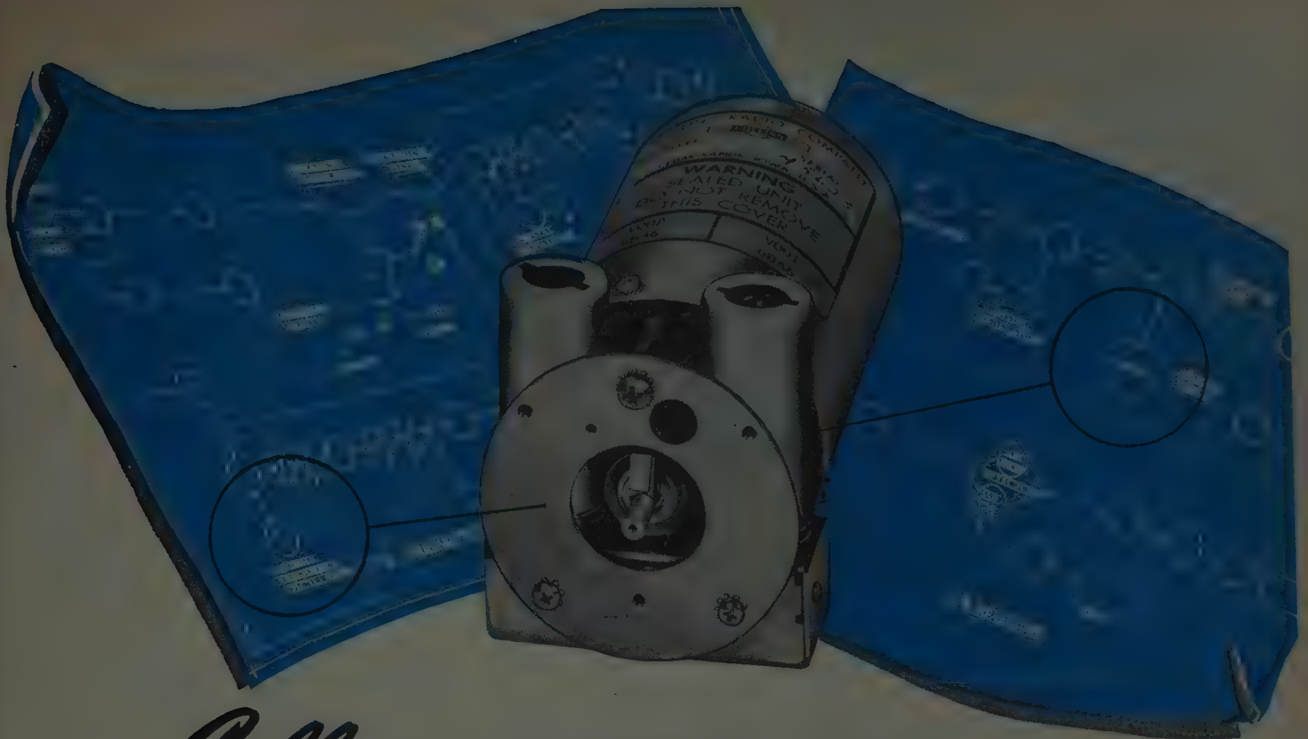
Membership

The following admissions and transfers were approved and are now effective:

Transfer to Senior Member

Aitken, K. M., Box 192, Merrit Island, Fla.
Alexander, S. N., National Bureau of Standards, Bldg. 10, Washington, D. C.
Atkins, G. T., 571 Nightingale Dr., Miami Springs, Fla.
Barus, C., Electrical Engineering Dept., Swarthmore College, Swarthmore, Pa.
Bowen, D. C., Radio Corp. of America, Bldg. 7-2, Camden 2, N. J.
Brewer, L. F., 168 Ludlow St., Portland 5, Maine
Buyer, E. M., 70 Grove St., Ramsey, N. J.
Camillo, C. C., 4358 S. Artesian Ave., Chicago 32, Ill.
Clayton, J. F., Bendix Aviation Corp., Research Laboratory Division, Bldg. 351, USNMATC, Pt. Mugu, Port Hueneme, Calif.
Earls, H. G., 202 Byron Rd., Fayetteville, N. Y.
Edgerly, J. J., Office of Naval Research, Special Devices Center, Port Washington, L. I., N. Y.
Fannin, B. M., 5206 Valley Oak, Austin, Tex.
Flemons, R. S., 1444 Sherwood Cres., Peterborough, Ont., Canada
Fragola, C. F., Sperry Gyroscope Co., Lake Success, L. I., N. Y.
Fraser, D. W., 2488B Morosgo Pl., N.E., Atlanta, Ga.
German, J. P., 1515 W. 32, Austin 3, Tex.
Green, P. E., Jr., 29 Granison Rd., Weston 93, Mass.
Haines, B. P., Philco Corporation, "C" & Tioga Sts., Philadelphia, Pa.
Hampton, J. W., 1627 Mayflower Dr., Irving, Tex.
Harpster, W. T., 48 Keats Ave., Town of Tonawanda, N. Y.
Hartwig, W. H., University of Texas, Austin, Tex.
Heaviside, M. G., 27 Whitney Rd., Newtonville 60, Mass.
Hogg, F. L., 37 Stormont Rd., Highgate, London N. 6, England
Houghton, E. G., Box 134, Hq. USAF Security Service, San Antonio, Tex.
Jackson, H. L., Westinghouse Electric Corp., 2519 Wilkens Ave., Baltimore 3, Md.
Jacobson, H. P., Jr., 1500 Oakland Rd., N.E., Cedar Rapids, Iowa
Junken, L. H., 118 Iroquois La., Liverpool, N. Y.
Kellogg, D. S., 9 Bradley Farms, Chappaqua, N. Y.
McCracken, L. G., Jr., 2540 Massachusetts Ave., N.W., Washington, D. C.
Mercier, A. E., 391 Fifth Ave., Cedarhurst, L. I., N. Y.
Meyerson, M., West Caranetta Terr., Lakewood, N. J.
Mickey, D. D., Jr., General Electric Co., 316 E. Ninth St., Owensboro, Ky.
Munn, A. J., Bell Telephone Laboratories, Inc., Whippany, N. J.
Palmer, J. R., 324 Homan Ave., State College, Pa.
Plotkin, M., Brookhaven National Laboratory, Upton, L. I., N. Y.
Ringoen, R. M., Collins Radio Co., Cedar Rapids, Iowa
Roth, J. H., 430 Princeton Rd., Haddonfield, N. J.
Ruehlmann, H. E., 1804 Jordan Park Apts., Fullerton, Pa.
Schauer, J. D., 408 Cornell, S.E., Albuquerque, N. Mex.
Shub, L., 299 University Ave., Kenmore 23, N. Y.
Smith, D. A., 505 S. Mesa Dr., Mesa, Ariz.
Waters, R. A., 4 Gordon St., Waltham 54, Mass.

(Continued on page 34A)



Collins PRECISION VFO—
Ready-to-Install ACCURACY and STABILITY

Accuracy and stability — the two most important features in Oscillator performance—can now easily be incorporated into your high-performance design, cutting engineering time to a minimum. Whether your project is a transmitter, receiver, test equipment, frequency standard or others, Collins offers a ready-to-install Variable Frequency Oscillator known for its linear calibration and stable output.

- Outstanding Stability
  - Average 24-hour stability under fixed-station conditions .003% or better.
  - Single-knob tuning with backlash of less than one cycle in 20 kc through use of mechanical loading and precision ballbearing construction.
  - Frequency modulation less than 100 cps under 5 G's acceleration at 60 cycles.
- Compact, ready-to-operate design.
- Linearity of calibration better than 1 kc throughout tuning range with multiple-turn tuning.
- Sealed against atmospheric changes.
- Available in fundamental ranges from 300 kc to 4 mc. Individual models achieve up to 2 to 1 tuning ratio.
- Uses standard power supply voltages.
- Each unit 100% tested under lab conditions to rigid specifications.
- Ease of installation.

Frequency Ranges Available

70E-1	1.0-1.5 mc
70E-10	600-800 kc
70E-12	1.955-2.955 mc
70E-15	2.0-3.0 mc
70E-20	1.65-2.05 mc
70E-21	300-400 kc
70E-25	2.0-4.0 mc
70H-2	2.455-3.455 mc
70H-3	1.5-3.0 mc

For requirements other than the above ranges or for detailed specifications write to the Collins office nearest you

COLLINS RADIO COMPANY CEDAR RAPIDS, IOWA

261 Madison Avenue, NEW YORK 16, NEW YORK
1200 18th Street N.W., WASHINGTON, D. C.
1930 Hi-Line Drive, DALLAS 2, TEXAS
2700 W. Olive Avenue, BURBANK, CALIFORNIA

COLLINS RADIO COMPANY OF CANADA, LTD.
77 Metcalfe Street, OTTAWA 4, ONTARIO

IRE Membership

(Continued from page 32A)

Wilde, A. E., Jr., R.D. 1, Zeek Rd., Morris Plains, N. J.
Williams, W. A., R.R. 3, Box 1033C, Evansville, Ind.

Admission to Senior Member

Albrecht, C. E., 2701 S. York St., Owensboro, Ky.
Bahls, W. E., Radio Corp. of America, 415 S. Fifth St., Harrison, N. J.
Callahan, J. G. P., 216 Telford Ave., Dayton 9, Ohio
Collins, F. V., Bell and Howell, Dept. 804, 7100 McCormick Blvd., Chicago 45, Ill.
Couillard, L. W., 2436 Franklin Ave., N.E., Cedar Rapids, Iowa
Dickey, W. S., 5663 E. 18 St., Tuscon, Ariz.
Ferguson, S. A., 632 Giralda Dr., Los Altos, Calif.
Froman, J., 1000 Geary St., San Francisco, Calif.
Fuerst, L. L., Edif. "Venus," Venus a Quebrada Honda, Caracas, Venezuela
Harries, W. L., 382 Marlborough St., Boston 15, Mass.
Jackson, W. R., Vickers Armstrongs Ltd., Weybridge, Surrey, England
Jones, H. S., Irco Corp., 16 Hudson St., New York 13, N. Y.
Lenzer, E., Fort Huachuca, Ariz.
Moore, N. H., 148 Tuscaloosa Ave., Atherton, Calif.
Nesbit, E. E., 15719 Rayen St., Supelveda, Calif.
Nicholas, J. C., Motorola Research Laboraotry, 3102 N. Ingleside Dr., Phoenix, Ariz.
Norton, C. A., c/o Remington Rand, Inc., Wilson Ave., S. Norwalk, Conn.
Patin, O. E., 12563 DeBell St., Pacoima, Calif.
Reeves, R., 96 Barons Ave., S., Hamilton, Ont., Canada
Taylor, C. A., Conestoga Farms, Conestoga Rd., Malvern R.D. 1, Pa.
Undesser, K., 3600 Bellevale Ave., Baltimore 6, Md.
Vine, B. H., 51 Northview Dr., Lancaster, Pa.
Vossen, B., 3903 N. Fourth St., Arlington 3, Va.

Transfer to Member

Abell, G. R., Jr., 82 College Ave., Poughkeepsie, N. Y.
Abel-Harry, C. B., 60 Highbury, London, England
Abrams, H., 123 Bragaw Ave., Newark 8, N. J.
Achatz, R. V., Southern Indiana Telephone Co., Aurora, Ind.
Achenbach, J. C., 50 Oak Ridge Dr., Haddonfield, N. J.
Adams, J. T., 408 Diamond Ave., Gaithersburg, Md.
Adams, M.M., 718 S. Delaware, Bartlesville, Okla.
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(Continued on page 36A)

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(Continued on page 40A)

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(Continued on page 42A)

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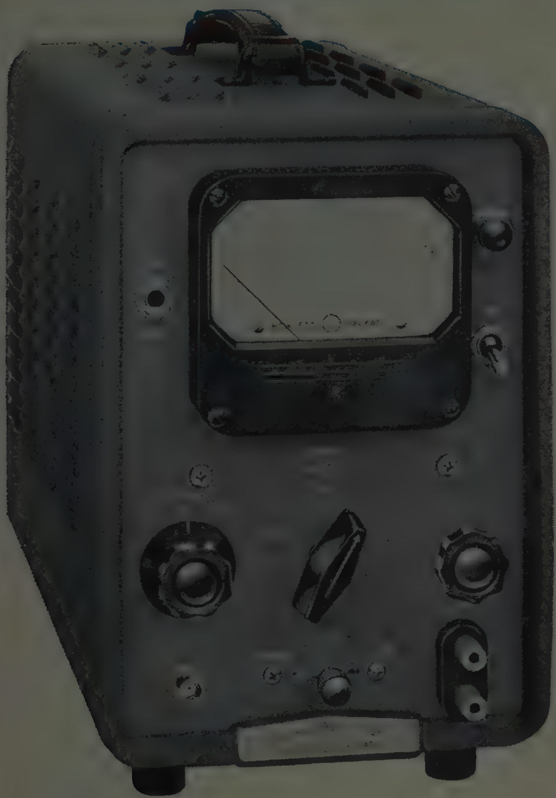

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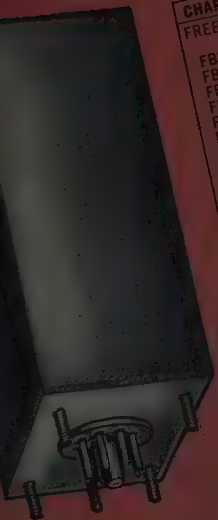
Covers the frequency band from 400 cps to 70 kc. 3 DB Band width ±9¾% of center frequency. Features octal "plug in" header. Size 2⅛" x 3½" x 4½".



CHARACTERISTIC IMPEDANCE 2500 OHMS

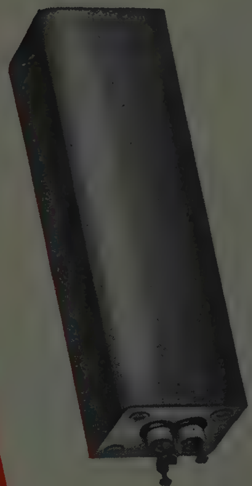
FREED #	CENTER FREQUENCY	3 DB BAND WIDTH
FBP-34	400 cps	± 9¾% CF
FBP 35	560 cps	± 9¾% CF
FBP 36	730 cps	± 9¾% CF
FBP-37	960 cps	± 9¾% CF
FBP-38	1300 cps	± 9¾% CF
FBP-39	1700 cps	± 9¾% CF
FBP-40	2300 cps	± 9¾% CF
FBP-41	3000 cps	± 9¾% CF
FBP-42	3900 cps	± 9¾% CF
FBP-43	5400 cps	± 9¾% CF
FBP-44	7350 cps	± 9¾% CF
FBP-45	10500 cps	± 9¾% CF
FBP-46	12300 cps	± 9¾% CF
FBP-47	14500 cps	± 9¾% CF
FBP-48	22 KC	± 9¾% CF
FBP-49	22 KC	±19½% CF
FBP 50	30 KC	± 9¾% CF
FBP-51	30 KC	±19½% CF
FBP-52	40 KC	± 9¾% CF
FBP-53	40 KC	±19½% CF
FBP 54	52.5 KC	± 9¾% CF
FBP-55	52.5 KC	±19½% CF
FBP-56	70 KC	± 9¾% CF
FBP 57	70 KC	±19½% CF

CHARACTERISTIC IMPEDANCE 500 OHMS

Band pass filters FBP-10 through FBP-33 feature the same attenuation characteristics as the above listing and are supplied with solder lug terminals.

DISCRIMINATOR INPUT LOW PASS FILTERS

Covers the frequency band of 400 cps to 70 kc. Less than .05 DB attenuation at ±9¾% of center frequency, 30 and 50 DB attenuation at the third and fifth harmonic of the pass band frequencies. Size 1½" x 1½" x 4½".



CHARACTERISTIC IMPEDANCE 30,000 OHMS

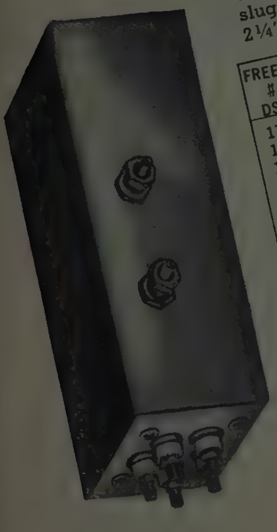
LPI —	Fo
10	400 CPS
11	560 CPS
12	730 CPS
13	960 CPS
14	1300 CPS
15	1700 CPS
16	2300 CPS
17	3000 CPS
18	3900 CPS
19	5400 CPS
20	7350 CPS
21	10.5 KC
22	12.3 KC
23	14.5 KC

CHARACTERISTIC IMPEDANCE 5100 OHMS

LPI —	Fo
24	22. KC
25	30 KC
26	40 KC
27	52.5 KC
28	70. KC

SLUG TUNED DISCRIMINATORS

Covers the frequency band of 3 kc to 70 kc. Frequency deviation ±8½% of center frequency. ½% linearity. Features slug tuned adjustment. Size 1⅝" x 2¼" x 4½".



FREED # DST	Fo	FREQUENCY DEVIATION	LINEARITY	DC OUTPUT
17	3 KC	± 8½% Fo	WITHIN .5%	32.5V
18	3.9 KC	± 8½% Fo	WITHIN .5%	32.5V
19	5.4 KC	± 8½% Fo	WITHIN .5%	32.5V
20	7.35 KC	± 8½% Fo	WITHIN .5%	32.5V
21	10.5 KC	± 8½% Fo	WITHIN .5%	32.5V
22	12.3 KC	± 8½% Fo	WITHIN .5%	32.5V
23	14.5 KC	± 8½% Fo	WITHIN .5%	32.5V
24	22 KC	± 8½% Fo	WITHIN .5%	32.5V
25	30 KC	± 8½% Fo	WITHIN .5%	32.5V
26	40 KC	± 8½% Fo	WITHIN .5%	32.5V
27	52.5 KC	± 8½% Fo	WITHIN .5%	32.5V
28	70 KC	± 7½% Fo	WITHIN .5%	32.5V
29	22 KC	±15% Fo	WITHIN 1%	26V
30	30 KC	±15% Fo	WITHIN 1%	26V
31	40 KC	±15% Fo	WITHIN 1%	26V
32	52.5 KC	±15% Fo	WITHIN 1%	26V
33	70 KC	±15% Fo	WITHIN 1%	26V
		FIXED DISCRIMINATORS		
10	400 cps	±8½% Fo	WITHIN .5%	32.5V
11	560 cps	±8½% Fo	WITHIN .5%	32.5V
12	730 cps	±8½% Fo	WITHIN .5%	32.5V
13	960 cps	±8½% Fo	WITHIN .5%	32.5V
14	1300 cps	±8½% Fo	WITHIN .5%	32.5V
15	1700 cps	±8½% Fo	WITHIN .5%	32.5V
16	2300 cps	±8½% Fo	WITHIN .5%	32.5V

DISCRIMINATOR OUTPUT LOW PASS FILTERS

Covers the frequency range from 6 cycles to 10,000 cycles. Features octal "plug in" header. Size 4" x 6" x 4½".

CHARACTERISTIC IMPEDANCE 330 OHMS

ATTENUATION: LESS THAN .2DB UP TO .5 TIMES Fo
LESS THAN .7DB FROM .5 Fo TO 1. Fo
MORE THAN 20DB AT 2 TIMES Fo
MORE THAN 30 DB FROM 3 TIMES Fo TO 100 KC

LPO—	Fo
10	6 CPS
11	8 CPS
12	11 CPS
13	14 CPS
14	20 CPS
15	25 CPS
16	35 CPS
17	45 CPS
18	60 CPS
19	81 CPS
20	110 CPS
21	160 CPS
22	185 CPS
23	220 CPS
24	330 CPS
25	450 CPS
26	600 CPS
27	660 CPS
28	790 CPS
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(Continued on page 50A)

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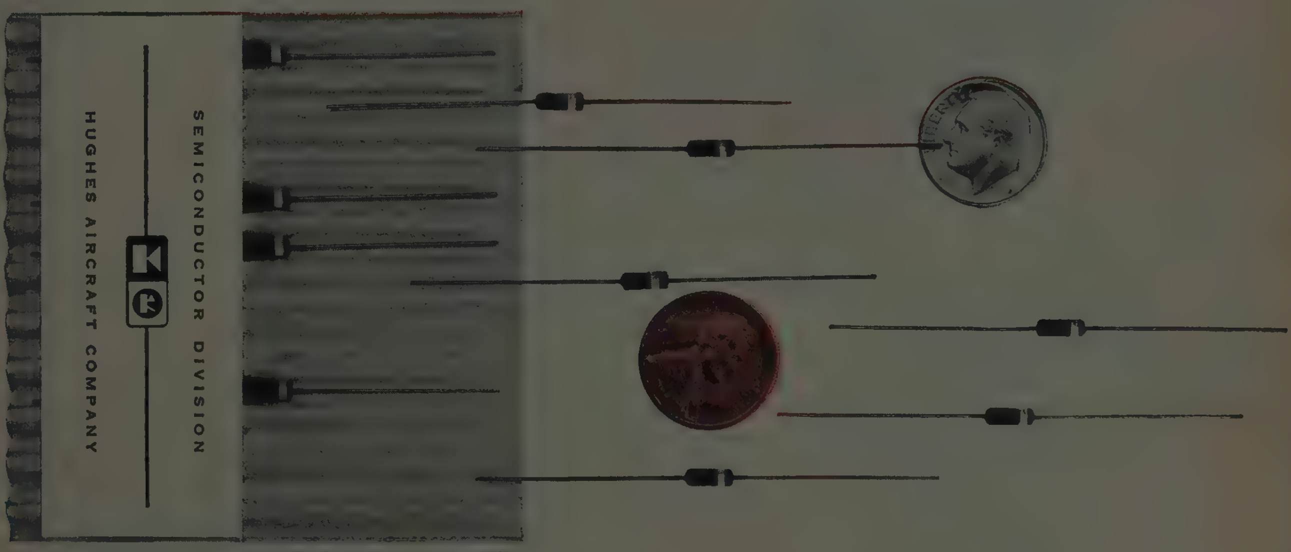

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(Continued on page 62A)

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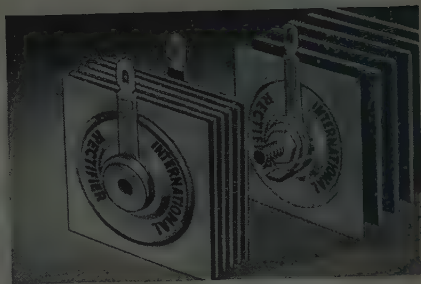

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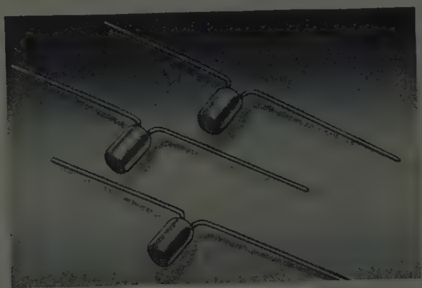

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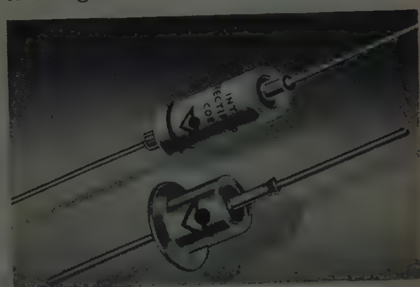

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(Continued on page 68A)

NEW high-power rectifier

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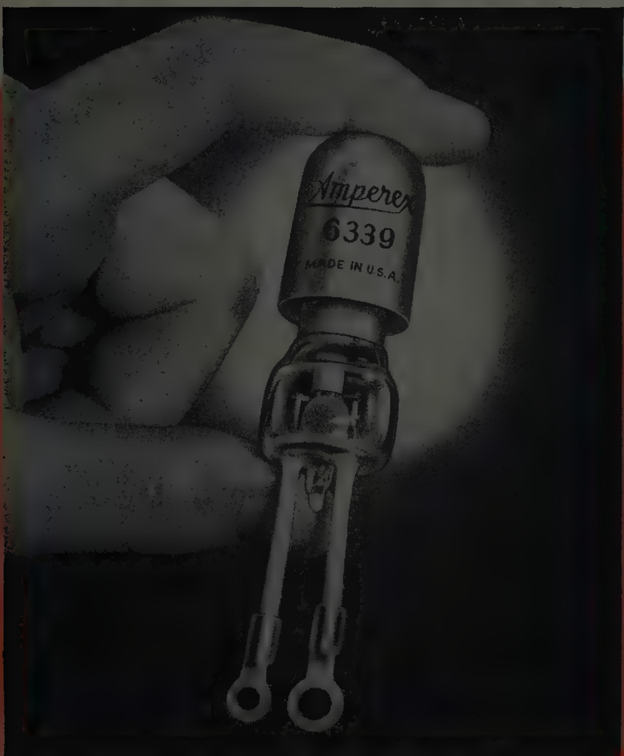

Only 2" (without leads) and ¾" in diameter, the AMPEREX 6339, a miniaturized, ruggedized version of the 3B29, operates under more stringent conditions than its prototype. This miniature, high-vacuum, external-anode, clipper diode and rectifier tube is designed to be enclosed in a complete liquid-cooled package, including power supply and pulse modulator components. It may also be operated in air at reduced ratings, in applications where liquid cooling is not required.

PARTIAL DATA — AMPEREX 6339

Filament Voltage	6.3
Filament Current	1.55

IN OIL — Maximum Ratings

Peak Inverse Voltage	10,000	16,000 volts
Peak Current	400	250 ma
Average Current	100	65 ma dc
Silicone Oil Coolant Temp.		−65°C to +165°C

Typical Operation (Three-phase, Bridge, Choke Input Filter)

No. of Tubes	6
Peak Inverse Voltage	16,000 volts
Peak Anode Current	195 ma
Average Anode Current (per tube)	65 ma dc
Output Voltage	14,000 volts dc
Output Current	195 ma dc

IN AIR — Maximum Ratings (sea level)

	Without Auxiliary Cooler	With Auxiliary Cooler
Peak Inverse Voltage	12,000	12,000 volts
Peak Current	200	400 ma
Average Current	50	100 ma dc
Ambient Temperature	−55 to +85	−55 to + 85°C

Typical Operation (Three-phase, Bridge, Choke Input Filter)

No. of Tubes	6	6
Peak Inverse Voltage	12,000	12,000 volts
Peak Anode Current	100	200 ma
Average Anode Current (per tube)	33	67 ma dc
Output Voltage	10,500	10,500 volts dc
Output Current	100	200 ma dc

The AMPEREX 6339 may be mounted in a standard 60 amp. fuse clip, as illustrated. For high-power operation in air, an auxiliary cooler which will also serve as a mount may be used. Flexible leads terminating in #6 and #8 lugs are used for heater and cathode connection. These allow the tube to be mounted in an extremely small space, and the leads may be brought out to any convenient terminal strip or stand-off insulator.

AMPEREX 6339 .... $35.00 net
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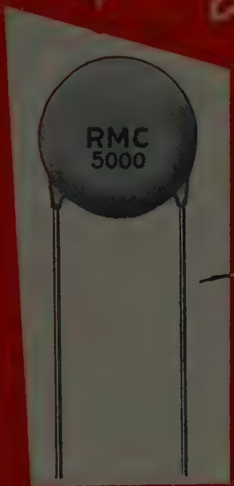

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(Continued on page 72A)

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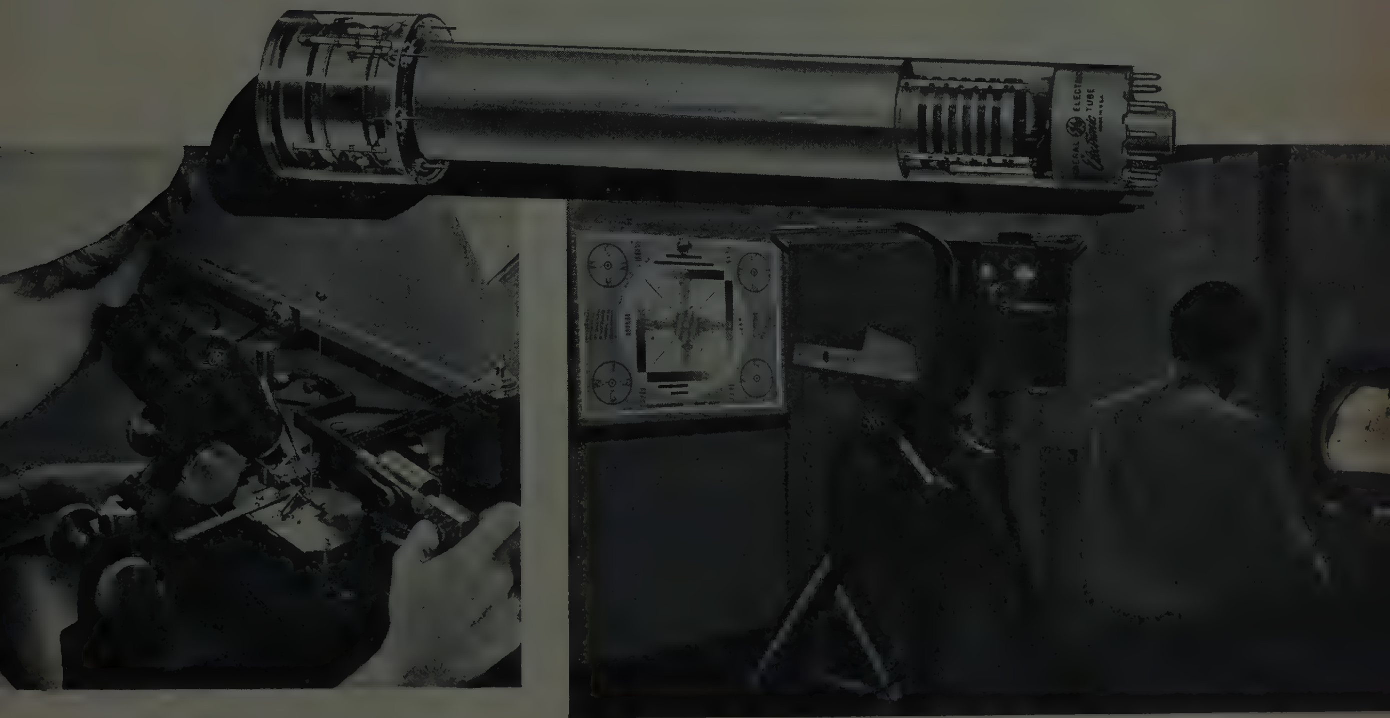

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(Continued on page 84A)

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(Continued from page 83A)

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(Continued on page 86A)

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(Continued from page 84A)

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(Continued on page 89A)

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(Continued from page 86A)

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(Continued on page 90A)

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(Continued from page 89A)

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(Continued on page 93A)

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(Continued from page 90A)

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(Continued on page 94A)

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(Continued from page 93A)

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(Continued on page 98A)

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CONTENTS

A Worthy Project

T. A. Hunter, Editor, IRE Student Quarterly

During the past few months I have had an opportunity to visit a number of sections actively participating in programs designed to be of service to our student members. Generally, these programs take the form of a student paper context, presentation of outstanding student awards, field trips, or career conferences. A recent innovation, undertaken by the Los Angeles Section, provides a student with the opportunity to discuss each area of the electronics industry, its future and the outstanding problems yet to be solved, with a representative of the Professional Groups. My experience with our student program indicates that this last type of program is by far the most effective and I wish to call it to your attention.

On May 3, 1955, the Los Angeles Section was host to four engineering colleges in Southern California with IRE Student Branches: California Institute of Technology, California State Polytechnic, University of California, and the University of Southern California. Mr. John O'Halloran, Student Branch Co-ordinator for the Los Angeles Section supervised the preparations. The student members, about 300 from the four colleges, were addressed by outstanding engineers, each representing one of the Professional Groups. Twenty minutes were allowed each group in addition to the open discussion which followed. A social hour concluded the afternoon program. Following the afternoon program students were allowed to discuss each group informally with the speakers in private rooms set aside for this purpose. During a dinner banquet following the conference Joseph Pettit, Regional Director, gave a short talk on IRE organization with emphasis on PG activities.

My own opinion of this meeting is certainly not as important as the opinion of the student body of the four schools. For this reason I have asked two students, Herbert Leach of California State Polytechnic College, and Paul Rude of the University of Southern California, to state their opinion of the meeting, together with those opinions they felt typified the views of the entire student body. Following are a few of their comments:

> "On May 3 the Los Angeles Section held a joint meeting with its student branches. The meeting was, we hope, the first of many. The purposes of the meeting were to promote the general professional development of the students through a better knowledge of the function of the IRE, to acquaint the students with leaders active in the IRE Professional Groups and to assist the students in selecting the phase of electronics in which they might wish to work after graduation.
>
> "The Los Angeles Section and its Student Relations Committee deserve much credit for a well-planned and very interesting day. The talks were presented in a manner so as to be of interest primarily to the students present. It

seemed that all of the speakers went out of their way to talk to the students and this consideration was quite effective.

"The afternoon session was sponsored by the various Professional Groups. The session was opened by a talk by Dr. Joseph Pettit of Stanford University, in which he discussed the attitude which the present IRE organization had towards the Professional Groups and how they have recently expanded. He and many of the following speakers emphasized the use of the Professional Groups by men in the field and the value of these groups in aiding members to keep abreast of the ever-changing electronics scene. Other speakers during this session spoke on their respective Professional Groups. These talks were of particular value to the students in that they dealt with basic philosophies in scientific and engineering investigations.

"Following dinner, recognition was paid to the chairmen of the student groups and student awards were presented to a student member from each of the student branches by Dr. Pettit. Following the presentation, Dr. Pettit spoke on 'What I Should Expect from the IRE,' giving a clear picture of the organization.

"During the regular section evening meeting Dr. Ernest Krause of Lockheed Missile Systems and Mr. John Byrne of Motorola spoke. Dr. Krause discussed the replacement of build and test, trial and error, experimental work in the field of missiles. Mr. Byrne spoke on recent developments in the field of mobile communication service.

"We feel, that as a whole, the meeting was very successful. We have made an attempt to contact some of the other students who attended the day-long session, and they seem to be in complete agreement with this opinion. Such meetings should be continued and held once each year. This type of meeting allows the student a chance to obtain a clear idea as to the organization and function of the IRE. Also students can meet with students from other colleges and broaden their views as to the various curricula and programs of the other student branches. Possibly in future meetings the students will be allowed to partake in the program itself. From the viewpoint of the Los Angeles Section the day should have been a success. Attendance was high and the response of the group was excellent. The section should take this hint and plan to continue such events."

I concur with both gentlemen and, based on my experience as editor of the *Student Quarterly*, I would further suggest that a recent graduate of an engineering college be invited to talk about his experience during his first year in industry. I have found that the readers of the *Student Quarterly* are most enthused about this type of material. I would also like to point out that the above response is truly typical of the many favorable comments brought to my attention—by the students themselves—during the conference.

On the basis of the enthusiasm shown it seems that the time and effort which went into making the program possible have been fully rewarded. As I left Los Angeles I felt pride in the fact that the IRE had members who would take the time to provide the student with such an unusual service. Any section wishing to undertake such a project is invited to write to Mr. John O'Halloran or me for a more complete description of the program. You will find the pattern one which is worthy of duplication.

Frequency and Time Standards*

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The following is one of a planned series of papers written at the invitation of the IRE, in which men of recognized standing review recent developments in, and the present status of, various fields in which noteworthy progress has been made.—*The Editor.*

Summary—Improvements in astronomical time measurement techniques and in the definition of time have kept pace with developments in frequency standards. Quartz crystal frequency standards are described, including Essen rings, bars, GT-cut plates, and contoured AT-cut plates. Stable oscillator circuits for quartz-crystal frequency standards are described, including the Meacham bridge-stabilized circuit, the Gouriet-Clapp circuit, and the Lea quartz-resonator-servo circuit. A discussion of the present status of atomic and molecular frequency standards includes the ammonia absorption cell, ammonia oscillator (MASER), and cesium atomic-beam apparatus. Instrumentation for precision frequency measurement is outlined, and a current listing of standard-frequency broadcast stations is included.

Introduction

Measurements of frequency and time have advanced in accuracy as the instrumentation for these measurements has improved. With each improvement in accuracy of measurement, new problems of stability, precision, calibration, and interpretation have become apparent. A review of the recent advances in frequency and time measurement technique is of interest to the radio engineer as an indication of the progress which has been made and of the improvements to be expected in the near future.

Time Measurement

The basis of frequency measurement is, axiomatically, time measurement, and conversely, time measurement can be based on frequency measurement. Before the discovery of atomic or molecular frequency standards, there were not available any alternatives to the calibration of frequency standards by means of astronomical observations. In view of the present early stage in the development of the atomic and molecular frequency standards, it is not yet possible to state that these atomic standards have been used to measure the constancy of astronomical time. However, the groundwork has been laid and soon it will be possible to calibrate astronomical time measurements against spectral-line frequencies. Further discussions of the spectral-line frequency standards are given in another section of this paper.

Accurate time is determined by astronomical observations at a designated observatory in each country where suitable observatories exist. The U. S. Naval Observatory is the only observatory in the United States regularly carrying out such measurements, and is thus the source of all accurate time determinations in this country. Time signals giving time as determined by the Naval Observatory are broadcast by naval radio stations,[1] and in cooperation with the Bureau of Standards, by stations WWV and WWVH which are operated by the Bureau of Standards.

A number of observatories in other countries are cooperating with agencies of their respective governments to furnish time measurements for radio transmission, and many of these observatories provide time measurements of very high accuracy. International comparison of time is carried on principally by means of radio transmission. (See section below on Standard Frequency Broadcasts.)

The problems of time measurement, and even of the definition of "time," have been familiar to the astronomer since long before the days of Sir Isaac Newton.[2] It is nevertheless true that our modern scientific notions of time are derived from the fact that time is the independent variable of Newtonian mechanics. Minor corrections, to take account of relativity, have enabled the original Newtonian concept of time to survive, and to provide a firm basis for astronomical time reckoning. As the stability of time-keeping devices has improved, it has become apparent that astronomers need to agree on a standard unit of time to use for astronomical calculations, and also to provide a basis for checking any variations in such time standards as the rotation of the earth. Consequently, in 1950 an international conference on astronomy recommended that the term Ephemeris Time be used to denote uniform or Newtonian time, and this term (Ephemeris Time) was adopted by the International Astronomical Union in September, 1952, as defining uniform time related to the revolution of the earth about the sun.[1,3] At the present writing,[4] it is impending that the International Committee on Weights and Measures will adopt a definition of the second, as a unit of time, as "the fraction 1/31,556,925.975

* Original manuscript received by the IRE, June 24, 1955.
† General Radio Co., Cambridge 39, Mass.

[1] Circular No. 49, U. S. Naval Observatory, Washington, D. C.; March 8, 1954.
[2] Dirk Brouwer, "The accurate measurement of time," *Physics Today*, vol. 4, pp. 6–15; August, 1951.
[3] Time Service Notice No. 1, U. S. Naval Observatory, Washington, D. C.; May 28, 1953.
[4] E. C. Crittenden, "International weights and measures, 1954." *Science*, vol. 120, p. 1008; December 17, 1954.

of the tropical year 1900." The adoption of this standard unit will serve to provide a time which may be used for data of great precision, such as may be required in frequency standardization.

In the preceding paragraph, the term Ephemeris Time was defined as denoting time based on the orbit of the earth around the sun. It is of interest to discuss the kinds of time and their significance in terms of astronomical phenomena. Ephemeris Time is determined by measurement of the tropical year. The tropical year is the time taken by the earth to make an orbit around the sun from vernal equinox to vernal equinox. By means of clocks, one can divide this tropical year into smaller intervals for application to various problems.

The time which is commonly used as "standard" time on the earth is determined by measuring the rotation of the earth about its own axis, especially with respect to the sun. Because of the ellipticity of the earth's orbit around the sun and the inclination of the earth's equator to the orbital plane, the length of an apparent solar day varies with the position of the earth on the ecliptic. In order to make the keeping of time independent of the seasons, the apparent solar day has been replaced by the "mean solar day," the duration of which is the average value of the apparent solar day over a period of a year. Very precise time measurements require corrections for the variation in longitude (apparent zenith) of the observing station and other small corrections known to astronomers.[5] Time determined by measuring the rotation of the earth was designated by the International Astronomical Union, September, 1952, as Universal Time. By international agreement, Universal Time is also defined as Greenwich Mean Time.

In order to provide a time measurement obtaining in one operation simultaneous data on the rotation of the earth and the rotation of a pair of bodies in space with a substantially constant rotational speed, observations of the moon and stars simultaneously have been undertaken.[6] The data obtained from such observations provides information on both Ephemeris Time and Universal Time, and it is thus possible to obtain an accurate difference term which enables precise conversion of one to the other.

It is expected that the above-mentioned improvements in observation techniques and method of computation of time will enable absolute frequency based on time measurements to be determined to approximately $\pm 1 \times 10^{-9}$.

The frequency of WWV, and of all standard frequency broadcast stations, is presently computed with respect to Universal Time (G.M.T.) which is mean solar time, thus automatically limiting the absolute accuracy to approximately $\pm 2 \times 10^{-8}$. This accuracy could be improved somewhat if corrections for short term variations in the earth's rotation were included in the computations.

Astronomical Time Measuring Instruments

Time determination requires specialized apparatus for the required astronomical observations. When visual observation is employed, the instrument most frequently used is the meridian transit telescope, which is constructed and mounted in such a way that it can be directed only at points along the meridian. The observer then operates the mechanism for recording the times of transit of the selected stars. Early designs of recording mechanisms depended on the reaction time of the observer to some extent. Improved designs have reduced the variation in observation from this cause, but the ultimate accuracy of measurement can only be reached when the observation can be made independent of the observer. Such independence from observer error can be achieved by photographic means, as in the apparatus described below.

Photographic Zenith Tube

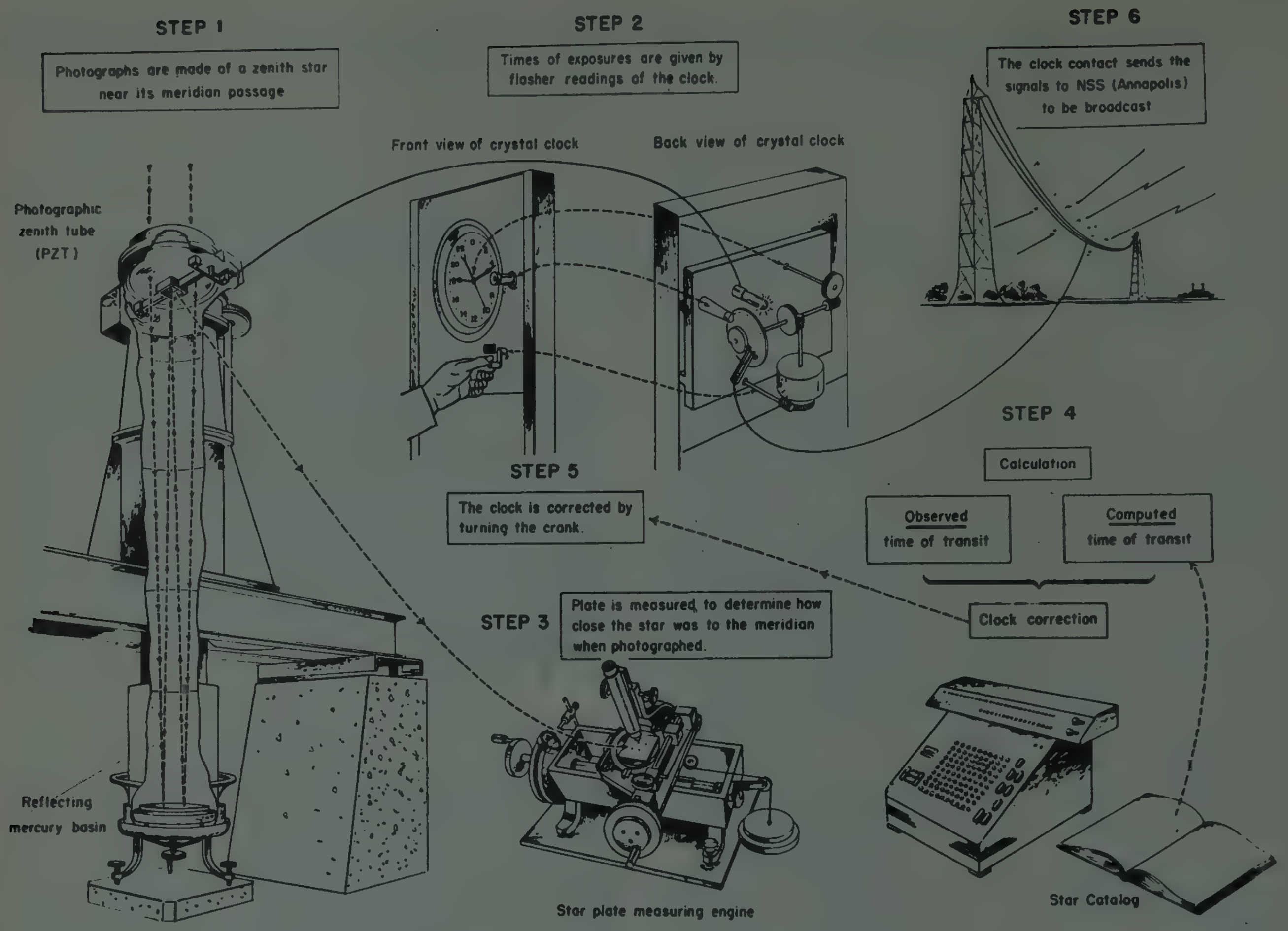
The principal device used by the U. S. Naval Observatory for the routine determination of star transits is the photographic zenith tube (PZT).[7] This device consists of a telescope of a special design for photographing stars near the zenith. A vertical tube is mounted above a mercury basin which, when used as a mirror, supplies automatically the vertical reference as a normal to its surface. The vertical or zenith view of this type of telescope minimizes the effects of atmospheric refraction and thus reduces observational errors. The upper end of the telescope tube supports the lens and the holder for the photographic plate used to record the positions of the stars. The plate holder is driven horizontally by an electric motor at a rate which synchronizes with the motion of the star images during two periods of exposure of the plate. Between the exposures, the carriage is rotated 180 degrees (images on opposite sides of the center of the photographic plate) thus providing simple and accurate geometrical determination of the meridian transit. The times at which the plate is at particular positions during the exposures are recorded on a chronograph driven by the crystal-controlled clocks of the Observatory. The positions of the stars are known, and it is thus possible to compute the correct time. An outline of the steps involved in the determination of time and transmission of time signals by the U. S. Naval Observatory is shown in Fig. 1 (next page). A photograph of a photographic zenith tube, PZT No. 3, at the U. S. Naval Observatory, is shown in Fig. 2 (page following).

Recent improvements in design of the plate carriage, motor drive, and chronographic pick-up of PZT have resulted in improved accuracy of time measurement.

[5] H. M. Smith, "The estimation of absolute frequency in 1950–51," *Proc. IEE*, vol. 99, pt. IV (Monographs), Monograph 39, pp. 273–278; December, 1952.

[6] W. Markowitz, "Photographic determination of the moon's position, and applications to the measurement of time, rotation of the earth, and geodesy," *Astron. Jour.*, vol. 59, pp. 69–73; March, 1954.

[7] W. Markowitz, paper on Photographic Zenith Tube now in preparation (U. S. Naval Observatory).



Official U. S. Navy photograph.

Fig. 1—Steps involved in the determination of time and transmission of time signals.

Dual-rate Moon Position Camera

Recently developed apparatus for observation and measurement of the position of the moon is now being applied to the problem of the precise measurement of time. The equipment and technique for obtaining a photograph of the moon simultaneously with that of the necessary stars for the calculation of the moon's position have been developed by W. Markowitz at the U. S. Naval Observatory.[6] The apparatus, or camera, for use on a refracting telescope comprises a special plate holder with a synchronous motor driving a micrometer screw to move the photographic plate at the sidereal rate corresponding to the moon's declination. The clock drive normally used to move the telescope tube is not used during this observation, the moving plate-holder being used instead. The image of the moon falls on a dark filter (attenuator) with a transmission factor of 0.001. This filter is a glass disk, with plane-parallel sides, 1.8 mm thick. Tilting of this disk about an axis parallel to its plane surfaces causes a translation of the image of the moon. A second synchronous-motor-and-micrometer drive controls the speed of tilt of this disk to hold the moon image fixed relative to the stars. A further adjustment enables selection of the axis about which the disk tilts. A chronograph contact registers the instant at which the photographic plate and the filter disk are parallel, i.e., the instant at which there is no relative shift in position between moon image and star images on the photographic plate. This instant thus defines the epoch of observation for time-measurement purposes.

A photograph of the dual-rate moon position camera installed on the 12-inch refractor of the U. S. Naval Observatory is shown in Fig. 3 (opposite page).

With the development of a satisfactory moon-star camera, it has now become feasible to institute a program of observation to chart the long-period variations in the rotation of the earth, and to compare them directly with Ephemeris Time determined from the same observations. (A group of photographic observations of moon and star positions was obtained at Harvard College Observatory, in 1911–17, and reduced at Princeton, but using another method.[6]) An extended series of such observations by several separated observatories is expected to be able to provide a basis for the determination of absolute frequency to 1 part in 10^9.

Frequency Standards

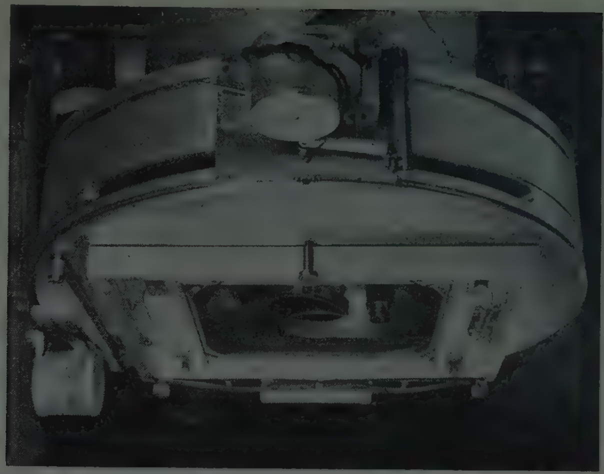
As may be inferred from the preceding discussion, the measurement of time by astronomical observation eventually requires extremely stable clocks in order to

Official U. S. Navy photograph.

Fig. 2—Photographic zenith tube, PZT No. 3, Naval Observatory, Washington, D. C.

provide means for subdividing a tropical year into 31,556,925.975 parts, each alike in duration. This extreme requirement for clock stability will be partially alleviated by the moon observation program which will provide monthly time checks. Clocks of the highest stability are necessary for scientific purposes such as the measurement of the short-period variations in the earth's rotation and the standardization of frequency.

The first crystal-controlled clock was constructed by W. A. Marrison and J. W. Horton in 1927.[8] Since that date, many engineers and scientists have made important improvements in the various components of the crystal-controlled clock, resulting in the stability mentioned above, and in impressive reliability as a laboratory tool for daily use, a reliability infrequently surpassed by any other electronic devices. Since the crystal clock is essentially a frequency standard with a cycle-counting device attached,[9] we shall here consider the various component parts of the crystal-controlled clock as being frequency standards and associated items, for it is as frequency standards that the radio engineer most often meets these elements of the crystal clock.



Courtesy *Sky and Telescope*

Fig. 3—Moon position camera, attached to the 12-inch refractor of the Naval Observatory.

Crystal-Controlled Frequency Standard Oscillators

In order to set forth the recent progress in frequency standard apparatus, it seems expedient to consider individually the elements making up such equipment. Most crystal-controlled frequency standards comprise (1) a control element, i.e., the quartz crystal unit, (2) a negative resistance element, i.e., the oscillator circuit using vacuum tubes or transistors to supply the power, (3) a thermostat or temperature-control device to keep the control element and other circuit elements at constant temperature, (4) suitable frequency dividers or other means for producing lower output frequencies, which may be used to operate (5) integrating devices, such as clock indicators, to keep a record of the number of cycles in a given period for comparison with astronomical time measurements. A suitable power supply (6) is, of course, required. Item (5) is sometimes eliminated in a secondary frequency standard if adequate reception is available from one or more of the standard frequency broadcasts now being transmitted by various agencies. Various other items of auxiliary equipment are frequently associated with crystal-controlled frequency standards for the purpose of calibration and standardization of the standards themselves, or for the use of the standards in frequency and time measurement.

Quartz Crystal Control Elements[10–12]

Two outstanding properties of crystalline quartz make it especially attractive as a control element for a piezo-electric oscillator, namely, the possibility of obtaining resonators of high Q value, and the exceedingly good stability of the quartz itself insofar as aging effects are concerned. Much of the frequency-standard work

[8] J. W. Horton and W. A. Marrison, "Precision determination of frequency," Proc. IRE, vol. 16, p. 137; February, 1928.

[9] W. A. Marrison, "The evolution of the quartz crystal clock," *Bell Sys. Tech. Jour.*, vol. 27, pp. 510–588; July, 1948. Also published as "Bell Telephone System Monograph B-1593," Bell Tel. Lab., New York City, and in *Horological Journal*, vol. 90, pp. 274 ff; May–October, 1948.

[10] R. A. Heising, "Quartz Crystals for Electrical Circuits," D. Van Nostrand Co., New York, N.Y., 1945.

[11] P. Vigoureux and C. F. Booth, "Quartz Vibrators," His Majesty's Stationery Office, London, England, 1950.

[12] J. P. Buchanan, "Handbook of Piezoelectric Crystals for Radio Equipment Designers," Wright Air Dev. Center (USAF) Tech. Rep. 54-248, Wright-Patterson AF Base, Ohio; December, 1954.

of recent years has been directed to the improvement of Q and aging characteristics of crystals.[13–15] The variation of frequency with temperature, an important matter for a stable oscillator, is a function of the shape of the crystal element, its dimensions and its angle of cut from the mother crystal. The pertinent properties of various types of crystal resonators currently considered suitable for use as frequency standards are considered herewith.

Rings and Bars

Crystal resonators operating in extensional modes offer some attractive properties for use at low frequencies. The choice of a suitable shape generally will provide one or more nodes suitable for use as mounting points, and the proper dimensioning, combined with a proper angle of cut, will produce a low coefficient of frequency vs temperature usually over a relatively narrow, specified temperature range. Such resonators at frequencies of the order of 100 kc have been made in the form of bars or rings.

Essen Ring[15]

A ring-type resonator, developed by Essen of the British National Physical Laboratory, has shown great stability in frequency-standard use. This resonator operates in the extensional mode with six half-wavelength sectors alternately extending and contracting in a direction along its circumference. The exciting voltage is applied to electrodes concentric with the inner and outer surfaces of the ring. Since the motion of the quartz is mainly along the circumference, there is only a little contraction and expansion of the surface of the ring and hence only a small power loss caused by ultrasonic radiation. An evacuated, sealed container has been used to keep the aging rate low, and incidentally also eliminate any residual losses caused by radiation from the ring or its mounting. The British-Post-Office Essen rings are reported to have a Q of two million,[16] while the earlier pin-type mount produced a Q of one million.

The Essen ring requires a fairly sophisticated mounting in order to take full advantage of its inherent high Q value. The mounting problem is simplified to some extent by the existence of the six nodal planes, which are zones of minimum vibration at 60 degrees angular separation around the ring. The earliest mountings made by Essen at N.P.L. employed pointed pins set into grooves cut into three of these nodal planes. Although the pins provided rugged support points, the rings seemed to exhibit some small frequency instability which was thought to be ascribable to the pin mountings. Consequently, a string or thread-type mounting was devised at the British Post Office for the Essen-ring crystal elements used in frequency standards designed there. Fig. 4 is a photograph of the Post Office Essen Ring, and

Courtesy H. M. Postmaster General.

Fig. 4—Photograph of 100 kc Z-cut quartz ring mounted on thread suspension in crystal holder W6, with cover removed.

Fig. 5 shows a sketch of the string mounting. The string-type mounting appears to have overcome the random frequency shifting observed with the pin-type support, but still leaves unsolved a few of the problems with respect to shipment or transportation of the finished quartz ring. The large mass of the Essen-ring crystal element imposes a requirement for relatively great care in shipment, requiring the type of shipment and handling normally reserved for delicate scientific instruments.

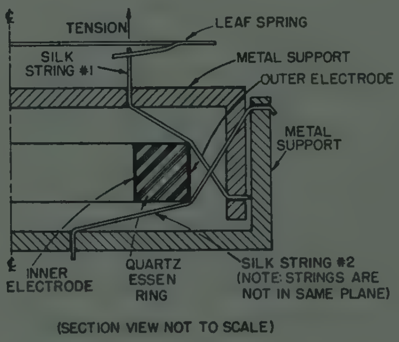


Fig. 5—Sketch of one of three string support points of Essen-ring crystal element.

Long term drift of the Essen-ring crystal is very small.[17,18] Values of drift rates of approximately 1×10^{-8} per month, or approximately 3×10^{-10} per day, have been observed for the Essen-ring oscillators at the U. S. Naval Observatory,[19] with the expectation that lower drift rates will be reached in the future. The lowest

[13] J. P. Griffin, "High-stability 100-kc crystal units for frequency standards," *Bell Lab. Rec.*, vol. 30, pp. 433–437; November, 1952.

[14] A. W. Warner, "High-frequency crystal units for primary frequency standards," Proc. IRE, vol. 40, pp. 1030–1033; September, 1952.

[15] L. Essen, "A new form of frequency and time standard," *Proc. Phys. Soc.* (London), vol. 50, p. 413; 1938.

[16] H. T. Mitchell and A. L. Dobbie, "100 Kc/s Oscillator of High Precision Incorporating an Essen Type Quartz Ring," paper presented at Congrès International de Chronometrie, Paris, France; October, 1954.

[17] H. M. Smith, "The determination of time and frequency," *Proc. IEE*, vol. 98, part II, pp. 143–153 (plus discussion); April, 1951.

[18] L. Essen, "Frequency standardization," *Proc. IEE*, vol. 98, part II, pp. 154–163 (plus discussion); April, 1951.

[19] Private communication.

drift rates of two such oscillators reported by the British Post Office are 0.25 and 0.4×10^{-10} per day over periods of several hundred days.[16] The British Post Office radio laboratory group considers that an Essen-ring oscillator unit is satisfactory for delivery to a user only if its drift rate is less than 5.0×10^{-10} per day averaged over 10 days. The excellent performance of the Essen ring with respect to long-term stability is ascribable, in part, to the fact that the frequency of oscillation of the ring is a function of the mean diameter of the ring, and that the loss, or acquisition, of a uniform layer of material over the entire surface would thus produce only a second-order change in the frequency. Careful processing of the ring and use of the evacuated mounting have further reduced the probability of changes in the crystal frequency.

An Essen ring ground for a frequency of 100 kc has an outside diameter of almost $2\frac{1}{2}$ inches (actually 61.26 mm in one case). This dimension is an indication of the difficulty of fabrication of such a crystal element, since it is necessary to obtain a quartz crystal free from defects with maximum dimensions large enough to allow cutting the ring from it. Because of this drastic requirement for large pieces of high-grade raw quartz-crystal, commercial Essen-ring frequency-standard units intended for moderate-quantity production have not been introduced.

Bars

Quartz bars vibrating in the extensional or longitudinal mode are widely used in frequency-standard oscillators. The attractive features of such bars include the availability of one or more nodal planes for the attachment of mountings, a large ratio of mass to surface for the finished crystal, and only a moderate size requirement for the raw quartz blank. In addition, the processing required is similar to that required for the more commonly used plates, i.e., plane lapping.

Frequency-standard crystals operating in the extensional mode have been used for many years. The German Physikalishe Technische Reichsanstalt group (Giebe, *et al.*) designed, constructed and operated for many years a quartz-controlled frequency standard using a 60-kc Y-cut bar.

A commercial frequency standard using a 50-kc X-cut bar was produced by the General Radio Company, Cambridge, Massachusetts, in 1928.[20]

A new design of overtone-operated X-cut bar was developed by Clapp[21] for use at 100 kc in the present model of the General Radio Company frequency standard (since 1947). This quartz bar (Type 1190-A Quartz Bar), shown in Fig. 6, operates at the second overtone, having two half-wavelength extensional mode sections operating in push-pull, i.e., the portion of the bar from the center to one end extends as the portion from the center to the other end contracts. A nylon-monofilament string suspension is used to support the bar at the two nodal planes, the filaments being maintained in tension by coil springs. Adjustable baffles at the ends of the bar are used to reflect ultrasonic radiation and thus reduce damping and change in frequency caused by changes in air pressure, as the mounting is not evacuated or hermetically sealed. Plated electrodes are applied directly to the surface of the bar on its sides, and are interconnected for second-overtone excitation in the extensional mode. The Q of this bar is approximately 170,000 in the mounting described.

Courtesy General Radio Company.

Fig. 6—Quartz bar for operation at 100 kc in second overtone mode. Note the end baffles to reduce ultrasonic radiation losses, and the string suspension at the two nodes.

Frequency stability of the commercial model bridge-stabilized oscillator, with which this bar is supplied in its temperature-controlled oven, reaches a value of approximately 0.5×10^{-8} per day or better, after an aging period of approximately one year. Many of these oscillators demonstrate considerably better stability than this figure. The long-term drift rate of the frequency standard in use at the General Radio Company has been approximately 5×10^{-7} per year since 1945, an aging rate of 1.2×10^{-9} per day averaged over 10 years.

Extensional-mode bars suitable for stable oscillator use have been made by other crystal manufacturers. Bars of the +5-degree X-cut, fundamental-mode longitudinal-vibration type, which were wire mounted with plated electrodes, have been used in a quartz-crystal-controlled clock in Switzerland.[22] These bars, mounted in evacuated glass envelopes, were supplied by Salford Electrical Instruments (British General Electric Company). They gave stabilities of the order of 0.5×10^{-8} per day, or better, when used in a Gouriet-Clapp oscillator circuit with automatic level control.

GT-Cut Plates

The GT-cut plate, originated by Mason,[23] has been developed to a highly advanced state for use in frequency standardization work.[18] This type of quartz

[20] L. M. Hull and J. K. Clapp, "A convenient method for referring secondary frequency standards to a standard time interval," Proc. IRE, vol. 17, pp. 252–271; February, 1929.

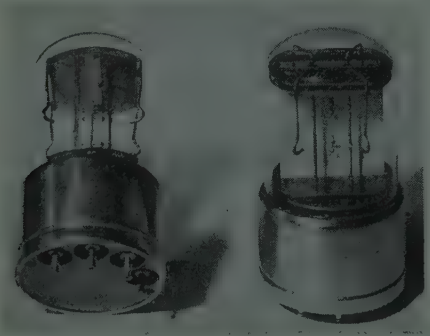
[21] J. K. Clapp, "On the equivalent circuit and performance of plated quartz bars," *Gen. Rad. Experimenter*, vol. XXII; March–April, 1948.

[22] P. Chalande, "The realization of a group of piezo-electric time-keepers," *La Suisse Horlogere* (International Edition in English), La Chaux-de-Fonds, Switzerland, pp. 41–44; October, 1952.

[23] W. P. Mason, "A new quartz crystal plate, designated the GT, which produces a very constant frequency over a wide temperature range," Proc. IRE, vol. 28, pp. 220–223; May, 1940.

plate can be made to have a temperature coefficient of frequency which is less than 2×10^{-7} per degree C., over a relatively wide temperature range. For the plates used as frequency standards, the temperature vs frequency curve is reasonably flat between 0 degrees and 100 degrees C., with optimum flatness in the range from approximately 20 degrees to 90 degrees C. Thus, the GT-cut plate can be made to serve as a stable element at temperatures approximating room temperature, and also at thermostatically controlled oven temperatures.

Early GT-cut plates were mounted on pressure-point contacts.[9,13,24] It is necessary to leave the edges of the GT-cut plate unsupported because they are vibrating with the greatest amplitude of any point on the crystal. Consequently, centrally located mounting points are desirable, the theoretical nodal point being at the center of the rectangular plate. Actually, because of couplings to other modes, the plates are not completely at rest at the central point. In addition, the desirability of keeping the attachment points small and flexible requires the use of several support points, which are now generally made in the form of thin wires attached to the surface of the crystal plate near the center, along the center line of the length of the plate.



Courtesy Bell Telephone Laboratories.

Fig. 7—100 kc GT-cut plates (D168670) in evacuated mountings as used in LORAN timer oscillators.

The use of GT-cut plates in frequency-standard oscillators was given impetus by the LORAN development during World War II which required stable oscillators for timing the pulses used in this radio-navigation system. Wire-mounted-silver-plated-electrode GT-cut plates were manufactured in evacuated glass envelopes for use in the LORAN timer oscillators. These crystals were a development of the Bell Telephone Laboratories, and represent an achievement of considerable magnitude in making a crystal unit largely independent of temperature, atmospheric changes, aging effects caused by exposure to the air, and a fair amount of rough handling in shipment. This crystal unit was designated by the number D-168670 (shown in Fig. 7, above).

Further refinement of this type of GT-cut plate has produced excellent results.[13] The improvements consist of reduction in the diameter of the support wires and their attachment points, improved methods of processing the soldered connections, and careful annealing to relieve strains. Final adjustment to frequency is accomplished by etching the edges. The electrodes are of gold to take advantage of the inherently stable character of this metal in this application. Twenty crystals were constructed for the National Bureau of Standards incorporating these improved design features, and are now in use by the Bureau of Standards at Boulder, Colorado, and at WWV.

The Q value of the D-168670 GT-cut crystals was approximately 140,000 and the frequency drift with time was approximately 1×10^{-8} per day in the LORAN oscillator. The Q value of the improved design is of the order of magnitude of a million, with some values as high as 4,000,000. The daily drift rate of the special GT-cut crystals in use at the National Bureau of Standards is reported as low as 1 to 5×10^{-10} per day, whereas the drift rate of the earlier design was reported as 1 to 3 parts in 10^9 per day after one year of aging.[13,25,26]

The principal advantage of the GT-cut appears to lie in its low temperature coefficient of frequency, and the consequent ability to provide a stable frequency even in the absence of precise temperature control. The National Bureau of Standards has demonstrated that it is possible to use a crystal resonator buried in the earth as a reasonably accurate frequency reference without further temperature control.[27] Such a system has the advantage that continuity of power supply is not necessary in order to preserve continuity in measurement of the aging curve of the crystal resonator, and that it is thus possible to use such a crystal as an emergency standard during a power failure.

It has been determined that GT-cut plates are sensitive to the amplitude of the driving current within the range of current experienced in the bridge-stabilized oscillator circuits normally used with these plates.[25] Although some improvement has resulted from redesign of the oscillator bridge networks to balance at lower values of crystal current, the National Bureau of Standards has incorporated into the group of crystals used as frequency standards several crystals which are used only as reference resonators; i.e., which are not running continuously in oscillator circuits but are measured in bridge circuits at low excitation current levels.[26]

[24] C. F. Booth and F. J. M. Laver, "A standard of frequency and its applications," *Jour. IEE*, vol. 93, part III, pp. 223–241 (with discussion); July, 1946.

[25] J. M. Shaull, "Adjustment of high-precision frequency and time standards," Proc. IRE, vol. 38, pp. 6–15; January, 1950.

[26] J. M. Shaull and J. H. Shoaf, "Precision quartz resonator frequency standards," Proc. IRE, vol. 42, pp. 1300–1306; August, 1954.

[27] T. A. Pendleton, "Underearth quartz crystal resonators," Proc. IRE, vol. 41, pp. 1612–1614; November, 1953.

AT-Cut Plates

The AT-cut quartz crystal plate was developed by Lack, Willard, and Fair in 1934.[28] Other investigators, notably I. Koga, also published data on similar low-temperature-coefficient cuts. This type of plate vibrates in the thickness-shear mode and may be made to have a low temperature coefficient of frequency. It is possible to orient the cut angle to produce an inflection point on the frequency-vs-temperature curve, that is, a zero temperature coefficient of frequency, in the range of temperatures normally used in temperature-controlled ovens. Such a crystal cut has obvious applications as a frequency standard.

Early efforts to use the AT-cut plates as standards[24] were hampered by the difficulty of mounting the plate in such a way as to achieve a mount which would not influence the frequency of the crystal. Low aging drift is almost impossible to attain unless a mount is used which affects the frequency of the crystal to a minimum degree. Booth of the British Post Office used nodal-plane pin-mounted AT-cut plates, operating at 1000 kc, with air-gap electrodes, in partially evacuated holders (air-pressure 3 cm Hg).[29] These crystals were operated at 50 degrees C. They gave drift rates averaging 2 to 5×10^{-9} per day over the years 1941–1944. In view of the fact that the nodal plane is in the center of the thin edges of the AT-cut plate (1.65 mm thick), the difficulty in constructing a stable mounting by this method was considerable.

The most promising recent development in the design of AT-cut plates for frequency-standard use has been carried out by Warner.[30] Warner has shown that a circular AT-cut plate with one side plane and the other side ground to spherical contour, operating at 5 mc in the 5th-overtone mode, can be made with a Q of approximately 2,500,000. A photograph of this crystal unit in an evacuated glass envelope is shown in Fig. 8. Warner further reports a 1 mc crystal of similar design[31] with a Q of 12×10^6. These remarkably high Q values are ascribable to the use of the overtone mode and to the spherical contouring, which "mismatches" the zones of the crystal away from the exact center of the convex side of the plate. The zones near the edge of the crystal are thus rendered incapable of resonant vibration at the excitation frequency and are consequently quiescent. The edge of the contoured plate is thus made suitable for the attachment of rugged mounting supports and connecting leads to the electrodes. The use of a glass-envelope evacuated mounting for this type of crystal plate has resulted in the high Q value quoted above, and in a low aging rate which is currently being verified at a number of laboratories. Indications are that the aging drift of this type of contoured AT-cut plate in an evacuated mount will be as low as that of any previously designed crystal units.

Courtesy Bell Telephone Laboratories.

Fig. 8—5 mc AT-cut contoured plate in evacuated glass envelope, operating in 5th overtone mode.

The advantages of the overtone-mode contoured 5 mc-plate for commercially produced equipment are centered in the relatively small size of the quartz blank required, and the ease of getting a satisfactory mounting. Careful processing is still necessary in order to attain low rates of frequency drift with time, but the ruggedness of the crystal unit and its small size have already suggested numerous applications.

Fundamental-mode AT-cut plates are capable of low rates of frequency change with time if properly processed and mounted, and if used in applications, such as high-stability circuits, where the constancy of the crystal can be exploited. Lea has used a fundamental-mode 5 mc plate in experimental oscillators of high stability,[32] and Sulzer has developed a 1 mc oscillator using a fundamental-mode contoured AT-cut plate.[33]

Oscillator Circuits for Frequency Standards

Resonant devices can be made to oscillate with good frequency stability only if appropriate means are selected for maintaining them in oscillation. Pendulum clocks furnish elegant illustration of this requirement.

[28] F. R. Lack, G. W. Willard, and I. E. Fair, "Some improvements in quartz crystal circuit elements," *Bell Sys. Tech. Jour.*, vol. 13, pp. 453–463; July, 1934.

[29] C. F. Booth, "The application and use of quartz crystals in telecommunications," *Jour. IEE*, vol. 88, part III, pp. 97–144 (with discussion); June, 1941.

[30] A. W. Warner, "High-frequency crystal units for primary frequency standards," Proc. IRE, vol. 40, pp. 1030–1033; September, 1952.

[31] A. W. Warner, "High-frequency crystal units for primary frequency standards," Proc. IRE, vol. 42, p. 1452; September, 1954.

[32] N. Lea, "Quartz resonator servo—a new frequency standard," *Marconi Rev.*, vol. 17, pp. 65–73; 3rd Quarter, 1954.

[33] "High-stability one-megacycle frequency standard," *NBS Tech. News Bull.*, vol. 38, pp. 162–163; November, 1954.

The Shortt clock, representing a highly developed form of the gravity pendulum clock using electrically supplied impulses to maintain oscillation, gives stability approaching that of crystal-controlled clocks. This stability is achieved by a combination of a stable resonator (free-pendulum), and an "oscillator circuit" which supplies a constant amount of power at the same point in every cycle. Similar requirements hold for quartz-crystal-controlled oscillators, each increase in stability of crystal elements calling for improvements in oscillator circuits.

The principal property susceptible to improvement is the stability of the phase shift in the "negative resistance" or amplifier element of the oscillator. There are currently at least three distinct approaches to the oscillator circuit problem, and possibly a great many more as yet not known to the author of this review. The first approach consists of the use of an amplifier with a positive feedback connection to provide regeneration and also frequency control through incorporation of the crystal element in this feedback path, with a negative feedback connection to stabilize amplifier gain and phase characteristics. The second approach comprises the use of the most stable elements in the "optimum" simple oscillator circuit with stabilization of the oscillator active element by appropriate means. The third approach adds to the second approach a servo-operated device for adjusting the circuit elements to maintain the oscillation frequency at a value which gives a constant value of impedance or phase shift in the crystal element.

Bridge-Stabilized Oscillators

The oscillator circuit which has been most widely used for frequency standard oscillators is the bridge-stabilized circuit originated by Meacham.[34] In this circuit [Fig. 9(a)], the feedback voltage which drives the amplifier is the unbalance voltage at the output terminals of a bridge network which includes the crystal with associated adjusting reactances, a resistor with a positive temperature coefficient of resistance, two linear-resistive arms and the necessary coupling circuits. The values of the resistors are so chosen, with respect to the crystal series resistance and the tungsten lamp resistance, that the bridge is unbalanced at low levels of applied signal in such a direction that positive feedback results from the bridge-unbalance output signal. As the amplitude of oscillation builds up, more current flows through the bridge arms, causing the tungsten lamp to increase its resistance and the bridge to approach the balance condition. The ultimate amplitude of oscillation is reached when the bridge unbalance signal becomes small enough so that the transmission loss through the bridge network equals the gain through the amplifier.

The excellence of the bridge-stabilized oscillator circuit stems from two important properties which are the result of the use of the bridge network in the feedback path. The first property is a function of the phase relationship of the input voltage of a bridge with respect to its unbalanced output, or detector output, voltage. Near the balance point of the Meacham bridge, incorporating the crystal resonator as one element, the slope of the phase shift of output voltage vs input voltage is greater than the slope of the phase shift of input voltage vs current through the crystal element alone. This improvement in slope enables design of oscillators in which improvement in stability is accomplished by provision of additional gain to make up for the loss involved in the operation of the bridge network close to the balance point. Improvement in frequency stability generally will result from increase in amplifier gain since the voltage gain goes up as the power of the number of stages, whereas the amplifier phase shift instability generally increases only directly with the number of stages.

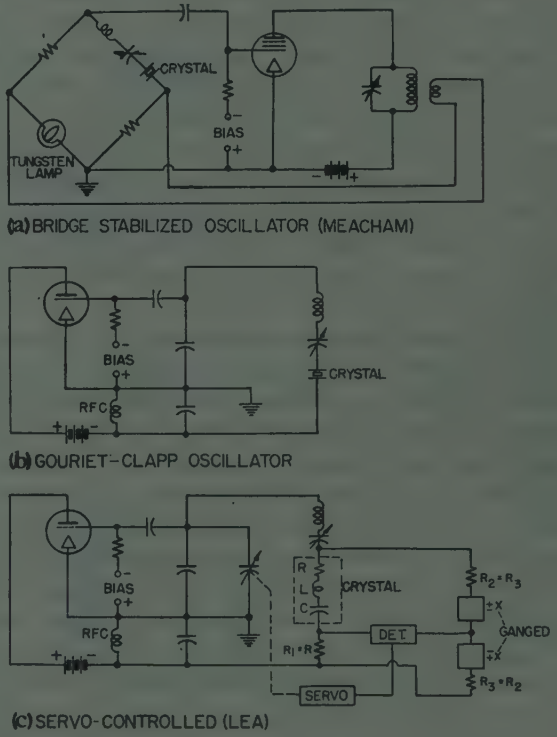


Fig. 9—Oscillator circuits for frequency standards.

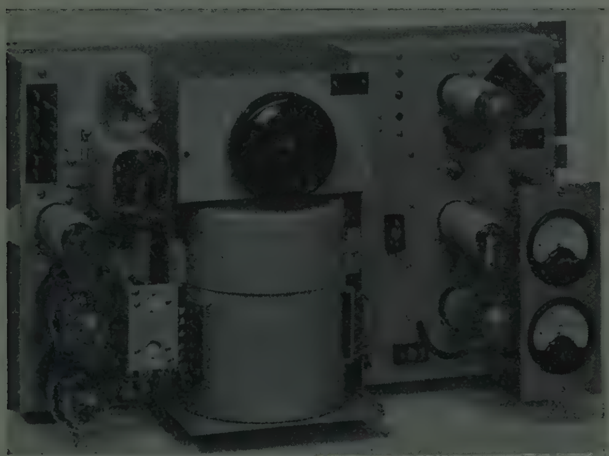
The second property of the bridge-stabilized oscillator, one which is at once an asset and a liability, is the amplitude stabilization property of the bridge network. The tungsten lamp has been almost universally used as the amplitude stabilizing device in the bridge because of its simplicity, ruggedness, and low drift with time. Some efforts have been made to use elements with negative temperature coefficients of resistance, such as thermistors, but the tungsten lamp is, in general, the accepted element. The amplitude stabilization resulting from the self-balancing bridge feedback network is effective, holding its amplitude setting well for long periods. However, the range of levels over which the bridge

[34] L. A. Meacham, "The bridge stabilized oscillator," Proc. IRE, vol. 26, pp. 1278–1294; October, 1938.

network can be made self-balancing, depends on the characteristics of the lamp, and generally higher levels are required than would be desirable for use with some crystal elements.[25]

The above-described properties of the bridge-stabilized oscillator are related to the general properties of amplifiers with feedback connections. It has been shown[35] that the performance of the bridge-stabilized crystal oscillator can be analyzed by separating the feedback circuit into a negative feedback path which stabilizes the gain and phase shift of the amplifier and a positive feedback path including the crystal unit, which determines the frequency of oscillation of the system. From this analysis, it appears that it may be profitable to explore further means for the stabilization of the amplifier circuits of oscillators.

Examples of the Meacham bridge-stabilized oscillator are provided by the LORAN timer oscillator (U. S. Navy, R. F. Oscillator Type 0-76/U),[36] the General Radio Company commercial frequency standard Type 1100-A, and the British Post Office Essen-ring oscillator, a photograph of which is shown in Fig. 10.



Courtesy H. M. Postmaster General.

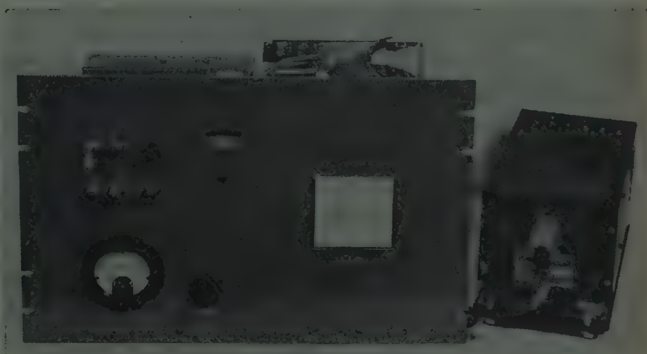
Fig. 10—British Post Office precision frequency standard oscillator, showing oven (center) containing 100 kc Essen ring. This oscillator uses the bridge-stabilized circuit.

Because of unavoidable stray inductance and capacitance, it has been generally found that the bridge-stabilized oscillator circuit is most useful at frequencies of 1 mc or below. Frequency-standard oscillators designed for operation at higher frequencies have, therefore, used the circuits described below.

Gouriet-Clapp Oscillator

The Gouriet-Clapp crystal oscillator circuit, shown in Fig. 9(b), has been used for many years in frequency monitors for broadcasting and in other applications where stable, simple oscillators are required. (This oscillator circuit is sometimes called a "modified Pierce" or "modified Colpitts" circuit. U. S. Patent No. 2,012,-497 was granted to J. K. Clapp for this crystal oscillator circuit in 1935, the series capacitance and inductance being adjusted to series resonance at the crystal series-resonant frequency. A similar circuit was developed independently by G. G. Gouriet of the B. B. C.) Recent availability of stable high-frequency crystals (See section on AT-Cut Plates, above) has prompted application of the Gouriet-Clapp circuit to frequency-standard oscillators in the megacycles/second range. An analysis (See Appendix) of the Gouriet-Clapp circuit with regard to the variations in frequency caused by changes in various circuit elements shows that an oscillator stability of the order of 1 or 2 parts in 10^9 should be realizable with this circuit using a crystal[30] with $Q=2.6\times10^6$ Application of automatic-gain-control to this oscillator circuit by controlling the grid bias of the vacuum tube with an amplified delayed-AGC circuit stabilizes the input impedance of the oscillator tube as well as the gain and crystal current.

Application of this circuit to frequency-standard oscillators has been carried out by Felch and Israel,[37] and in considerably modified form, by Lea (See Servo-Controlled Oscillators, below). The stability achieved has been 3×10^{-9} per day or better, using the 5 mc overtone-mode AT-cut plate,[30] by the former group. A photograph of this 5 mc oscillator unit is shown in Fig. 11.



Courtesy Bell Telephone Laboratories.

Fig. 11—Photograph of USAF Type 0-269 (XW-1)/UR Oscillator, using 5 mc overtone-mode contoured AT-cut plate (see Fig. 8).

Servo-Controlled Crystal Oscillators

All of the oscillator circuits described above have relied on the steep slope of phase-change with frequency in the crystal element to provide corrections for the drifts of phase in the oscillator circuit in order to maintain a constant frequency of oscillation. The bridge-stabilized oscillator alone has provided an enhanced phase-change system to assist the phase-vs-frequency

[35] E. J. Post and H. F. Pit, "Alternate ways in the analysis of a feedback oscillator and its application," Proc. IRE, vol. 39, pp. 169–174; February, 1951.

[36] J. A. Pierce, A. A. McKenzie, and R. H. Woodward, "Loran," McGraw-Hill Book Co., New York, N. Y. ("Model UE-1 Oscillator," pp. 237–240, describes the Type 0-76/U Oscillator); 1948.

[37] E. P. Felch and J. O. Israel, "A simple circuit for frequency standards employing overtone crystals," Proc. IRE, vol. 43, pp. 596–603; May, 1955.

slope of the crystal element. An oscillator circuit which provides a somewhat different method of frequency control has been developed by Lea.[32] A simplified circuit diagram of his servo-controlled oscillator is shown in Fig. 9(c). It should be noted that the servomechanism has been added to an oscillator circuit which, for purposes of illustration, is similar to the Gouriet-Clapp oscillator of Fig. 9(b). The short-term or cycle-to-cycle phase stability of the oscillator is thus dependent on the Q of the crystal, which Q has been degraded to $\frac{1}{2}$ of its original value by the addition of $R_1=R$ (crystal series resistance). The phase "noise" or phase instability of this circuit may thus be twice that of the Gouriet-Clapp circuit using the same crystal unit. However, the long-term stability (for any period longer than the correction time of the servo control) is determined by the ability of the servo system to maintain the oscillator frequency at that value which appears to result in a constant value of impedance in the crystal unit. A bridge circuit comprising the crystal (R, L, C) R_1, R_2, R_3 and ganged modulating reactances $\pm X$ and $\mp X$ is provided by adding $R_1=R$ in series with the crystal, and adding R_2, $\pm X$, $\mp X$ and R_3 in parallel with the crystal branch. R_2 should be equal to R_3, but be large compared with R (crystal) and R_1. A detector, comprising a sensitive AM receiver, is provided with a phase-detector output circuit synchronized with the modulation rate of $\pm X$ and $\mp X$. If the frequency applied to the crystal deviates from the frequency of the crystal series resonance, the voltage drop across the crystal arm of the bridge will change both in magnitude and in phase. The modulating reactances, $\pm X$ and $\mp X$, being modulated continuously at a fairly constant rate, will enable sensing of the direction of phase change of the bridge unbalance voltage (detector output) by scanning back and forth through a small range of reactance unbalance in the modulating arms, and using a phase-sensitive circuit tuned to the modulation frequency at the output of the detector. The output signal from this detector will then be proportional to the magnitude of the deviation from bridge balance, and will have a phase or sign which indicates the direction of deviation of the applied frequency from the crystal resonant frequency. The detector output signal is then applied to a servo system to readjust the oscillator circuit to reduce the frequency deviation to a minimum. By increasing the gain of the detector circuit, it is possible to reduce the magnitude of the deviation required to operate the servo device until the limiting signal-to-noise ratio is reached.

The servo system is thus used to correct for such instability as may arise in the "negative resistance," that is, in the vacuum tube (or transistor) and associated reactive elements. Instability is thought to arise from such factors as cathode-interface impedance, space-charge capacitance, changes in tube geometry with age, transit time variation, and perhaps Miller-effect capacitance changes in addition. A delayed-automatic-gain-control is used by Lea to stabilize level and grid input impedance. The correction time of the servo control used is fairly short, a variable capacitor being driven by a motor to effect the adjustment of the circuit reactance. In its present state of development,[38] the servo-controlled oscillator is stable to better than $\pm 3\times 10^{-11}$, and the average frequency to approximately $\pm 1\times 10^{-11}$, for periods in excess of 10 seconds, the ultimate drift rate for long periods thus being dependent only on the constancy of the crystal element except for the $\pm 3\times 10^{-11}$ error of the circuit. This figure includes changes of tubes, drift of the feedback circuit elements, and supply voltage changes.

Further application of this servo-control principle has produced comparable results using slightly different circuit details. Lea makes use of a motor-driven variable inductance as a single modulated reactance, dispensing with the second modulated element, and delayed AGC. Sulzer[39] has used a chopper to commutate small capacitors in the modulated reactance positions, and a limiter to control level. Both systems operate at a modulation rate different from the power frequency in order to avoid "hum" troubles.

Tuning Forks as Frequency Standards

Tuning forks have been used as frequency standards.[24] The advantages of the tuning fork as a clock-driving source derive mainly from the low frequency of oscillation of the fork and the simplified auxiliary apparatus needed to drive the clock. Interest in small, lightweight, frequency standards for airborne applications has kept the tuning fork from being completely eclipsed. Several manufacturers are producing hermetically-sealed temperature-compensated tuning forks operating in the frequency range of 400 to 1,000 cps, and also at 50–60 cps, and at some frequencies above 1 kc. Performance of the best of these tuning-fork units is comparable with that of commercial-grade crystals as far as stability is concerned. For example, one of these forks (at the Riverbank Laboratories, Geneva, Illinois), operating without temperature control at room temperature in an amplitude-stabilized oscillator circuit, has given stability of the order of $\pm 1\times 10^{-6}$ for several weeks. The Q realizable in a tuning fork is limited, and consequently, the instantaneous phase stability of an oscillator circuit using fork control has to be made as high as possible in order to keep the frequency from fluctuating rapidly. With a modern tuning-fork-controlled oscillator, it is possible to realize a portable time and frequency standard with stability adequate for many purposes.

Microwave Spectral Lines of Atoms and Molecules as Frequency Standards

Much has been written of the many proposals for the use of the constant properties of atoms and molecules as standards of frequency. It will not be possible here to

[38] Private communication, February 1, 1955.
[39] P. G. Sulzer (National Bureau of Standards), "High stability bridge-balancing oscillator," paper in preparation.

give a complete description of the status of the various projects in this field of endeavor, but the projects which appear most promising will be covered briefly. Of the many possible spectrum lines in the microwave region, the 3,3 inversion-line of the ammonia molecule (NH_3) and the transition (4,0 3,0) of the cesium atom seem to be nearest to practical application. Both of these spectrum lines have already been used as the bases of frequency calibrating apparatus,[40] and it is probable that their use will result in the first frequency standards of high precision with complete freedom from long-term aging drift. If present theories of atomic structure are rigorously correct, and there appears to be no reason for suspecting otherwise, then the frequencies representing the spectral lines should never change. We should, therefore, be able to use these invariant frequencies as frequency standards without reference to astronomical phenomena except for initial calibration. It is probable that the first frequency and time standardization using these spectral lines as standards will be done by using them as calibration standards to measure the constancy of the frequency of a conventional frequency standard or of the oscillator of a quartz-crystal-controlled clock, and thus enable accurate establishment of the time-keeping rate of the clock for comparison with astronomical time. As the perfection of atomic frequency standards progresses, it may prove feasible to use them as standard-frequency oscillators for routine laboratory measurements.

The problem then resolves itself into the design of equipment and the application of the information obtained from the equipment. Since the techniques for the two spectrum lines mentioned above are so widely different, they will be treated individually.

Ammonia Spectrum Line Developments

It is probable that the earliest published reference to the possibility of using microwave spectrum lines as frequency-stabilizing elements is in a paper by Pound[41] published in 1946, although other investigators had perceived the possibility of using the microwave spectral lines as frequency calibration points. Shortly after publication of Pound's paper, a paper by Smith, de Quevedo, Carter and Bennett[42] confirmed the application of Pound's method of stabilization using the 3,3 line of ammonia (NH_3) as the frequency reference. The stabilized oscillator system comprised a reflex klystron, a wave-guide hybrid system, a wave-guide resonator filled with ammonia, and a "dc" feedback connection to the klystron repeller electrode to close the loop. In effect, the ammonia was used as a resonant element to provide a rapid change of phase of a reflected wave in a Pound-type discriminator, the rate of change of phase with frequency being rapid enough to give an effective Q estimated at 12,500.

The use of the 3,3 inversion line of ammonia at approximately 23,870 mc for this stabilization experiment was the extension of many years of investigation of this particular spectrum line. Cleeton and Williams measured this ammonia absorption in 1934,[43] and a number of papers appeared immediately after World War II[44–46] giving further information which indicated that the 3,3 line of ammonia was a strong line (high absorption of energy), and that it was not affected in frequency by such variable factors as pressure, temperature, and magnetic field, although the apparent resolution or breadth of the line depends on pressure and temperature.

The most accurate determination of the frequency of the 3,3 inversion line of ammonia appears to be that by Shimoda,[47] who gives a value of $23{,}870{,}130.97 \pm 0.10 \pm 1$ kc for this line. This figure includes terms of ± 0.10 kc instrumental error, and ± 1 kc uncertainty concerning the absolute value of the reference frequency standard.

Servo-Controlled Ammonia Oscillators

A method of oscillator stabilization using a control loop and an ammonia absorption cell as a frequency-stable element has been applied to frequency-standard oscillators. Hershberger and Norton[48] stabilized a klystron oscillator at the ammonia-line frequency, and also offset from this frequency by a known intermediate frequency increment. Lyons[40,49] applied a similar approach to the stabilization of a crystal-controlled frequency-standard oscillator, and thus to the control of a clock by reference to the ammonia absorption-line frequency. Fletcher and Cooke stabilized a klystron at the ammonia-line frequency.[50]

The basic principles of such a servo-controlled oscillator are shown in Fig. 12(a) (next page). An oscillator, with a controllable frequency adjustment, supplies a signal to a modulation system which adds modulation to the signal, which is then referred to the ammonia-filled absorption cell. The signal is modified by passage through the cell, the modification then being detected and evaluated by the circuits of the servo control with

[40] H. Lyons, "Spectral lines as frequency standards," *Ann. N. Y. Acad. Sci.*, vol. 55, pp. 831–871; November, 1952.
[41] R. V. Pound, "Electronic stabilization of microwave oscillators," *Rev. Sci. Instr.*, vol. 17, p. 490; November, 1946.
[42] W. V. Smith, J. L. G. de Quevedo, R. L. Carter, and W. S. Bennett, "Frequency stabilization of microwave oscillators by spectrum lines," *Jour. Appl. Phys.*, vol. 18, p. 1112; December, 1947.
[43] C. E. Cleeton and N. H. Williams, "Electromagnetic waves of 1.1 cm wavelength and the absorption spectrum of ammonia," *Phys. Rev.*, vol. 45, pp. 234–237; February 15, 1934.
[44] C. H. Townes, "The ammonia spectrum and line shapes near 1.25 cm wavelength," *Phys. Rev.*, vol. 70, p. 665; November, 1946.
[45] W. E. Good, "The inversion spectrum of ammonia," *Phys. Rev.*, vol. 69, p. 539; May, 1946.
[46] B. Bleaney and R. P. Penrose, "Ammonia spectrum in the 1 cm wavelength region," *Nature*, vol. 157, p. 339; May, 1946.
[47] K. Shimoda "Atomic clocks and frequency standards on an ammonia line," *Jour. Phys. Soc. Japan*; Part III, 1954.
[48] W. D. Hershberger and L. E. Norton, "Frequency stabilization with microwave spectral lines," *RCA Rev.*, vol. 9, pp. 38–49; March, 1948.
[49] H. Lyons, "The atomic clock, an atomic standard of frequency and time," *NBS Tech. News Bull.*, vol. 33, pp. 17–24; February, 1949.
[50] E. W. Fletcher and S. P. Cooke, "The stabilization of a microwave oscillator with an ammonia absorption line reference," Cruft Laboratory, Harvard University, Tech. Report No. 5; 1948, Tech. Report No. 64; 1950.

reference to the modulation system. The servo control then supplies a correction to adjust the frequency of the controlled oscillator to the desired value.

Hershberger and Norton[48] swept the frequency of a separate klystron local oscillator back and forth across the frequency of the ammonia cell, and detected the pulse resulting from the absorption peak. Simultaneously, they applied this FM signal to a mixer with a signal from the controlled oscillator (a reflex klystron) and amplified the beat-notes near zero-beat (pulses) resulting from this interaction. The phase of the two sets of pulses was compared, and a correction signal obtained which was contrived to move the controlled oscillator pulse to coincidence with that from the ammonia cell. A further arrangement was constructed which used an offset, or intermediate-frequency, beat-note from the controlled-oscillator part of the circuit to provide the control pulses. By using a stabilized intermediate frequency, a stable controlled frequency resulted.

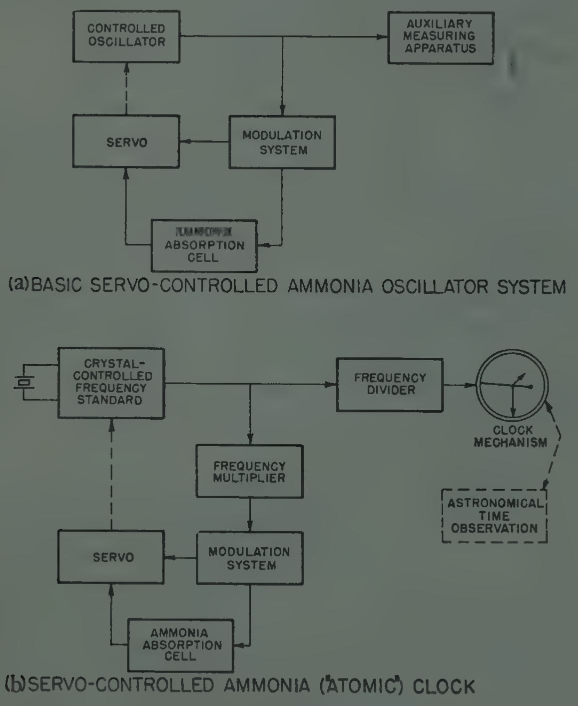


Fig. 12—Servo-controlled ammonia-absorption-cell oscillator systems.

The atomic clock development program under Lyons at the Bureau of Standards has explored the possibility of stabilizing a crystal-controlled frequency standard against the ammonia absorption cell.[40,49] The ammonia-stabilized clock uses a system of stabilization similar to the one discussed above [see Fig. 12(b)], but resulting in a lower output frequency which can be used to operate a clock mechanism for comparison with astronomical time measurements. The controlled oscillator feeds a frequency multiplier chain which eventually provides output near the frequency of the ammonia line. At one stage in the multiplier system, a frequency-modulated signal is added to that from the multiplier stage, and the proper sideband signal selected to provide a harmonic falling on the 23,870 mc frequency of the ammonia cell. Thus it is possible to provide a frequency-modulated signal derived from the frequency standard, sweeping back and forth in the vicinity of the ammonia frequency, with good short-term stability of the center (or carrier) frequency. The intermediate-frequency frequency-modulated signal (that which was added to the multiplied frequency of the controlled oscillator) is compared with the appropriate harmonic of the controlled frequency-standard oscillator, a signal pulse being produced each time the swept intermediate-frequency signal passes a given reference frequency. The FM signal, at 23,870 mc $\pm$modulation, undergoes absorption each time it sweeps past the ammonia absorption frequency in the cell, this absorption being observed as a negative reference pulse out of the detector at the receiving end of the ammonia absorption cell. The servo circuits are operated by the phase or time difference between these two pulses and are arranged to produce a correction of the crystal oscillator frequency to keep the crystal-controlled frequency standard locked to the ammonia line. The result which is sought is to produce a clock with no net long-term drift in its time-keeping rate, and with good short term stability, or low acceleration. A clock constructed on these principles gave a performance estimated at $\pm 2\times 10^{-8}$ for a period of the order of one week. The average frequency or integrated time error was not determined. A photograph of the first ammonia clock built at the National Bureau of Standards (1948–1949) is shown in Fig. 13 (opposite). The ammonia absorption cell is mounted in a coil around the large clock indicator above the racks.

A different approach to the servo-control system problem was used by Fletcher and Cooke.[50] Their modulation system used frequency modulation of the controlled oscillator at a relatively high modulation frequency but with a low modulation index. This modulation produced two sidebands which were on either side of the frequency range affected by the ammonia absorption line. An amplitude-modulation detector was used at the output of the absorption cell. If the phase of the carrier (23,870 mc) of the FM oscillator became shifted from its original phase by the action of the ammonia absorption, amplitude modulation resulted upon recombination with the unshifted sidebands.[51] This amplitude modulation occurred at the modulation frequency of the FM ("intermediate frequency"), the AM signal being recovered by the AM detector at the receiving end of the absorption cell. This intermediate frequency signal was then amplified and compared in phase with the modulating signal, the output of the phase comparison circuit being applied to the repeller electrode of the controlled klystron oscillator as a dc adjustment of the average frequency of oscillation.

[51] M. G. Crosby, "Communication by phase modulation," Proc. IRE, vol. 27, pp. 126–136; February, 1939.

Courtesy *Annals of New York Academy of Sciences*

Fig. 13—Photograph of first ammonia clock built at National Bureau of Standards.

A stabilized oscillator using an ammonia absorption cell modulated by a Stark-effect modulator was constructed by Townes[52] in 1951.

Difficulties in the Use of Ammonia Absorption to Stabilize Oscillators

Certain basic difficulties beset the use of the ammonia absorption technique for the stabilization of oscillators.[40,52] The principal difficulties of an inherent nature (properties of molecules) are (1) the natural breadth of the spectral line, (2) Doppler-effect broadening, (3) pressure broadening caused by collisions between molecules, (4) broadening caused by collisions with the walls of the absorption cell, and (5) saturation effects. The natural line breadth is related to the radiation from the molecule and the amount of thermal radiation falling on it. It is inherent and cannot be changed except by choice of the molecule or atom to be used. The other effects are usually much greater, in any case. Doppler-effect broadening is proportional to the velocity of the gas molecules parallel to the propagation direction of the radio-frequency energy in the cell. It can be reduced by cooling, but the ammonia freezes[46] out if cooled far enough to provide much reduction. Pressure broadening results because the energy absorption process is interrupted if a molecule collides with another during the absorption, and has to start again with a new phase possible. This effect can amount to 15 mc bandwidth at a pressure of 1 mm of Hg, but it diminishes with pressure reduction. Wall collisions cause broadening, but amount to a relatively minor item, of approximately 15 kc bandwidth maximum. Saturation effects result from the possibility of all available molecules having already been excited to the higher energy state. and those which are emitting energy supplying enough quanta to re-excite those which require excitation. The only energy then absorbed at the inversion line frequency is that lost to thermal radiation by collision and radiation damping of the moledules. The power input level to the absorption cell at which saturation effects set in is proportional to the square of the pressure in the cell, and hence is conflicting with pressure broadening effects as far as the selection of a pressure level for the cell is concerned.

In addition to the theoretical limitations set forth in the preceding paragraph, the design and construction of the microwave rf system for an ammonia-absorption-cell stabilized oscillator is complicated by the difficulties of working in the frequency range close to 23,870 mc. The signal-to-noise ratio of the system is affected by the noise in the detector, in particular, and could be improved if the saturation effects did not limit the allowable power input. The design of a cell to hold the ammonia gas is complicated by the necessity for maintaining a low standing-wave ratio over the band of frequencies used by the modulation system chosen. A schematic showing the principal features of one design of ammonia absorption cell is shown in Fig. 14 (next page), and a photograph of an ammonia absorption cell is shown in Fig. 15 (next page).

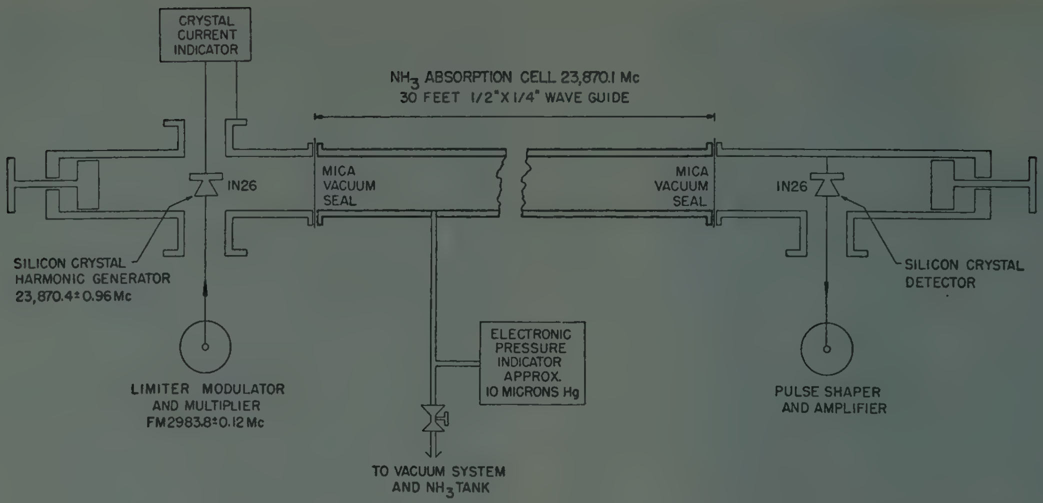
Further work on the solution of these problems appears unlikely in the future as a result of the success of other approaches to the atomic-frequency-standard problem, although experimental work on absorption cells will undoubtedly continue.

Ammonia Oscillator

A completely different arrangement for the use of the 3,3 inversion-line of NH_3 as a frequency standard has been devised by Townes,[53] of the Department of Physics, Columbia University, New York City. Ammonia gas at room temperature contains molecules in various energy states. Slightly less than half of the molecules are in the upper-energy states, while the remaining molecules are in the lower states. The lower-energy-state molecules have an electric dipole moment which makes it possible to accelerate them in a given direction by putting them in an electric-field gradient. The molecules in the upper states are accelerated in the opposite direction along this same electric-field gradient. Thus a sorting or selecting device may be constructed by setting up an appropriately shaped transverse-electric-field gradient in a region traversed by a stream of ammonia molecules, the lower-energy-state molecules being diverged away

[52] C. H. Townes, "Atomic clocks and frequency stabilization on microwave spectral lines," *Jour. Appl. Phys.*, vol. 22, pp. 1365–1372; November, 1951.

[53] J. P. Gordon, H. J. Zeiger, and C. H. Townes, "Molecular microwave oscillator and new hyperfine structure in the microwave spectrum of NH_3," *Phys. Rev.*, vol. 95, pp. 282–284; July 1, 1954.



Courtesy *Annals of New York Academy of Sciences*

Fig. 14—Diagram of ammonia absorption cell for atomic clock.

Courtesy *Annals of New York Academy of Sciences*

Fig. 15—Photograph of ammonia absorption cell for atomic clock.

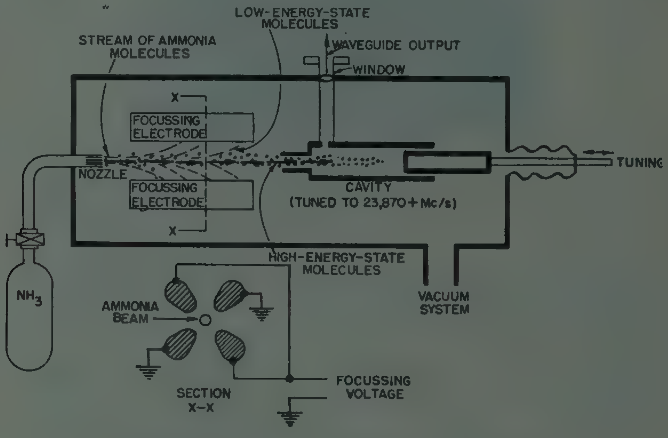


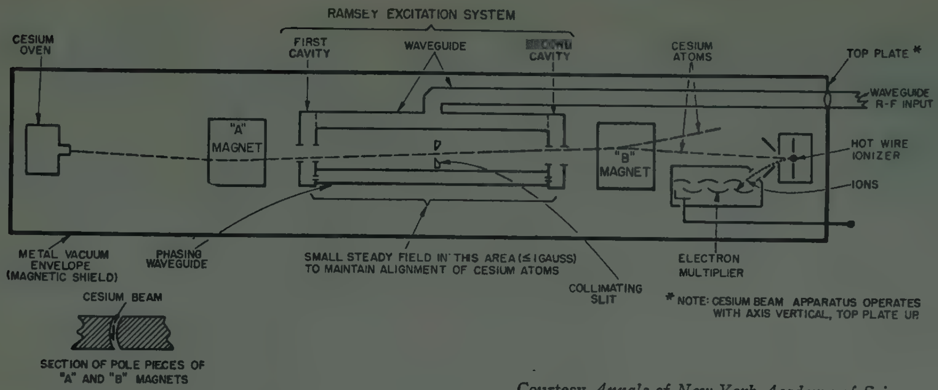
Fig. 16—Ammonia oscillator (Townes).

from the axis, and the higher-energy-state molecules converged by the focusing system. By this means, a useful portion of the high-energy molecules in a given stream may be selected and focused at the end of the electrode system.

Such a system is shown schematically in Fig. 16, with a resonant cavity to receive the focused high-energy molecules through a waveguide-below-cutoff entrance port. This device operates to produce oscillations at the inversion-line frequency of the ammonia by the following mechanism: the high-energy-state molecules which enter the resonant cavity are acted upon by any radio-frequency fields present in the cavity, and also these molecules can contribute to these field components by emission of energy. Some of the molecules in the cavity undergo transition to the lower state by emission of a quantum of energy at 23,870 mc. When the rf field at this frequency builds up to a sufficient value, the transitions are stimulated and the molecules then give up their quanta in an ordered, coherent manner, thus providing a source of power at 23,870 mc. The magnitude of the power available is adequate to supply the losses in the radio-frequency circuit, and to provide an additional small amount of power for measurement purposes (estimated 10^{-8} to 10^{-9} watt).

The general class of devices of this sort has been designated MASER, from the initials of the description "*m*icrowave *a*mplifier by *s*timulation of *e*mitted *r*adiation." In the case discussed above, the gain of the amplifier is greater than the losses in the system, and hence an oscillator is the result.

The exact frequency at which the oscillations are produced depends on several factors, the two most significant ones being the Q of the cavity and the tuning of the cavity relative to the inversion-line frequency. In the first experimental models of this device, detuning



Courtesy *Annals of New York Academy of Sciences*

Fig. 17—Cesium atomic-beam frequency-standard apparatus.

the cavity produced a "pulling" effect of approximately $\pm 2{,}000$ cycles. At the present time, the best method of estimating the correct center frequency of the 3,3 inversion line of ammonia appears to be setting to the midpoint of the pulling range. Other methods may be devised with better reproducibility of setting, such as the use of the frequency at which oscillations are just observable when the Q of the system (cavity plus load) is reduced to the point where self-oscillations are barely possible.

Two of these oscillators are reported to have been operated simultaneously, beating one against the other in a receiver tuned to their frequency. The oscillators were detuned to produce a 50-cps beat note, and the instability was observed to be less than ± 0.1 cps. Over a period of an hour, the average variation in the beat-note was less than ± 2.5 cps and the peak deviation was less than 5 cps.

It is estimated that a fully engineered version of this type of oscillator may reach a long-term stability of $\pm 1 \times 10^{-12}$. The absolute accuracy of the oscillation frequency cannot now be specified, but it is apparent that the oscillator may be set by simple methods to within approximately ± 20 cps of the correct frequency, or $\pm 1 \times 10^{-9}$, and that improvements in setting techniques will improve this figure.

Cesium Atomic-Beam Frequency Standard[40,54,55]

Another atomic spectrum line which may be used for frequency standardization is the line at 9,192.63197+ mc which is observed in cesium of atomic weight 133 by atomic-beam techniques. The atomic or molecular beam apparatus for measuring nuclear magnetic moments by resonant absorption was developed by Rabi and coworkers at Columbia University.[56] The original laboratory equipment gave a minimum indication upon the absorption of a quantum of any frequency, whereas the present models give a maximum indication upon the absorption of a quantum at the desired frequency only. The energy level difference corresponding to this frequency in the cesium atom is associated with the spin vector of the valence electron and its relation to the nuclear magnetic moment of the atom, the two energy levels corresponding to the case of the electron spin vector being aligned with and in the same direction as the nuclear magnetic moment, and the case in which the spin vector is directly opposed to that of the nucleus. When an atom of cesium is acted upon by a magnetic field of exactly the correct frequency, the internal structure of the atom can absorb a quantum of energy corresponding to the transition described above. The external evidence of this change in energy level is provided by a change in the magnetic moment of the atom. The atomic-beam apparatus shown in Fig. 17 is designed to enable detection of the changed magnetic moment of the atoms, and hence to determine the correctness of the frequency of the exciting field in the cavities. The width of the resonance curve of the absorption line is inversely proportional to the time the atom spends in the exciting field, the time in this case, using the two-cavity excitation method, being the time taken to traverse the path from the entrance of the first cavity to the exit from the second cavity.

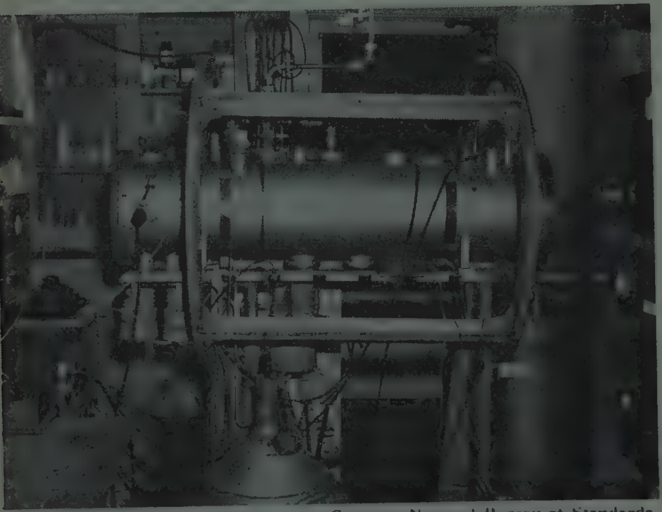
The cesium-beam apparatus shown in the diagram (Fig. 17) is typical of current designs. A stream or beam of cesium atoms is emitted by the oven through a nozzle which provides a ribbon-shaped beam of approximately 0.02-inch thickness, the emission of cesium being approximately 10^{-6} grams per day. The atoms pass through the inhomogeneous magnetic field between the polepieces of the A magnet. Those atoms with the appropriate dipole moment are deflected by the magnetic field gradient of the A magnet as indicated in the diagram, and are turned back toward the axis of the apparatus. The cesium atoms then traverse the first rf cavity in which they are exposed to a magnetic field at 9,192+ mc which can produce the energy level change desired in the atoms. The atoms then drift through the

[54] J. R. Zacharias and J. G. Yates, "VIII, Atomic Beam Research; A Cesium Clock," Quarterly Progress Report, Research Laboratory of Electronics, Mass. Inst. Tech., Cambridge, Mass., pp. 30–34; October 15, 1954.

[55] N. F. Ramsey, "Nuclear Moments," John Wiley and Sons, Inc., New York, N. Y., ch. 3, sec. D, "Molecular Beam Resonance Methods," pp. 37–52 (An extensive bibliography is given on this general type of molecular beam apparatus); 1953.

[56] I. I. Rabi, S. Millman, P. Kusch, and J. R. Zacharias, "The molecular beam resonance method for measuring nuclear magnetic moments," *Phys. Rev.*, vol. 55, pp. 526–535; March 15, 1939.

distance between the cavities (50 to 100 cm) and then through the second rf cavity. The radio-frequency magnetic field in the cavities is set to the same phase by careful adjustment and is checked by means of a probe inserted in the phasing waveguide connecting the two cavities. The net effect of the use of two separate in-phase cavities is similar to the effect obtained by using a long cavity with zero phase-shift between the ends, with the exception that, at frequencies slightly separated from the center of the resonance curve, an interference pattern occurs which shows up as a large amplitude ripple in the main absorption curve. This method of excitation, originated by Ramsey,[57] provides a sharper peak at the center of the resonance curve than is provided by the use of a single excitation field, reducing as it does the Doppler effect to a very small value. The atoms which have absorbed (or emitted) a quantum in the space between the magnets have then changed their magnetic dipole moment and are deflected in the opposite direction by the magnetic field gradient in the B magnet, while those atoms which have not "flopped" are deflected a second time, as before, and are not refocused on the detection device. The detection device comprises a surface ionizer, of the hot wire type, which is hit by the neutral atoms, and ionizes them. The cesium ions thus formed are then accelerated and focused by the appropriate electrodes and injected into the secondary-emission electron multiplier. The output current of the electron multiplier collector electrode is thus a measure of the number of atoms making the transition, and hence of the resonance curve of the transition.



Courtesy National Bureau of Standards.

Fig. 18—The National Bureau of Standards cesium atomic-beam equipment.

Fig. 18 shows a photograph of the atomic-beam portion of the cesium-beam frequency-standard apparatus constructed at the National Bureau of Standards.[40] The path length between the rf cavities is 50 cm. The effective Q obtained was 30 million. The atomic beam was horizontal in this apparatus. The excitation for the rf system is supplied through the waveguide entering the top of the container. Control of the ambient magnetic field affecting the equipment is provided by the large coils surrounding the vacuum envelope. The crystal-controlled excitation system is not shown in this photograph.

Current practice makes use of a small amount of frequency modulation of the exciting oscillator and appropriate phase-sensitive circuits to control the average frequency of the exciting oscillator. Hence the cesium-beam apparatus is a form of servo-controlled oscillator with a highly specialized form of absorption cell in which the Doppler effect is very small, collision broadening is absent, and which uses a very sensitive, low-noise, detection circuit not heavily limited by saturation or detector thermal noise level.

The excitation oscillator used in such a system must be adequately stable in order to avoid spurious effects, and the auxiliary equipment associated with the atomic-beam apparatus requires careful design in order to provide the best stability and accuracy for the over-all frequency-standard apparatus. The excitation system used at the Bureau of Standards is crystal-controlled at a relatively low frequency and uses a multiplier chain to reach the operating frequency of the cesium beam. Another suitable oscillator system has been constructed at M.I.T., using a Western Electric Type 416-B microwave triode working at approximately 3,064 mc and tripling with crystal diodes.

The cesium-beam apparatus which has been run at M.I.T. is reported to have a stability of $\pm 1 \times 10^{-9}$ for short periods, with the mean frequency showing less drift than this value. Refinements in this apparatus are expected to improve the over-all stability. A newer projected design is also being undertaken in an effort to improve the over-all performance by several orders of magnitude.

A commercial model of the cesium-beam atomic frequency standard is now being designed (by the National Company, Malden, Massachusetts) and should be available shortly.

Further Atomic Frequency Standards

Although the spectrum lines of atoms and molecules in the microwave frequency range are almost limitless in number, only a few of these spectrum lines offer attractions comparable with those of the lines described above. Dicke is carrying out work at Princeton which may result in the use of a line of the sodium spectrum as a reference. Some frequency calibration measurements on oxygen absorption lines are being carried out at the National Bureau of Standards. However, at the

[57] N. F. Ramsey, "A molecular beam resonance method with separate oscillating fields," *Phys. Rev.*, vol. 78, pp. 695–699; June 15, 1950.

TABLE I

PRINCIPAL CHARACTERISTICS OF STANDARD-FREQUENCY AND TIME-SIGNAL STATIONS

Stations	Hawaii	Johannesburg[5]	Rugby	Tokyo	Torino	Uccle[22]	Washington
Call-sign	WWVH	ZUO	MSF	JJY	IBF	—	WWV
Service	Experim'l	Experim'l	Experim'l	Experim'l	Experim'l	Experim'l	Regular
Carrier Power (kW)	2[1]	0.1	0.5	1	0.3	0.02	10[1]
Type of antenna	Vertical dipole	Inverted L	Vertical dipole	Vertical dipole	Horizontal dipole[18]		Vertical dipole
Number of simultaneous transmissions	3	1	3	1	1	1	6
Number of frequencies used	3	1	3	3	1	1	6
Transmission							
Days per week	7	7	7	7–2[12]	1[19]	7	7
Hours per day	22	24[6]	24[9]	24	6[20]	22	24
Standard frequencies used							
Carriers (mc)	5, 10, 15	5	2.5, 5, 10[10]	2.5,[13] 5,[14] 10[15,16]	5	2.5	all
Modulations (cs)	1,[2] 440, 600	1[7]	1,[2] 1,000	1,[17] 1,000	1,[2] 440, 1,000	None	1,[2] 440, 600
Duration of tone modulation (minutes)	4 in every 5[3]	—	5 in every 15	9 in every 20	5 in every 10[21]	—	4 in every 5[3]
Accuracy of frequencies (10^{-8})	±2	±2[8]	±2	±2	±2	±1	±2
Max. oscillator drift (10^{-8}) per month	+2	+4	+0.5	+1	+4	—	+1
Max. value of steps of frequency adjustment (10^{-8})	1	2	2	2	2	—	1
Duration of time signals in minutes	continuous	continuous	5 in every 15	continuous	5 in every 10	None	continuous
Accuracy of time intervals	$\pm 2\times 10^{-8}$ ± 1 µs	$\pm 2\times 10^{-8}$ ± 10µs	$\pm 2\times 10^{-8}$ ± 1 µs	$\pm 2\times 10^{-8}$ ± 1 µs	$\pm 2\times 10^{-8}$ ± 1 µs	—	$\pm 2\times 10^{-8}$ ± 1 µs
Method of adjusting time signals	Steering[4]	Steering[4]	By steps of 50 ms[2]	Adjusted to mean of time signals	Steering[4]	—	Steering[4]

[1] Maximum values, reduced power is used on certain frequencies and on certain days, [2] 5 cycles of 1,000 cps modulation pulses, [3] 440 and 600 cps alternately, [4] No phase adjustment to the signals themselves, [5] Transmission by the Union Observatory (Union of South Africa), [6] Interruptions for short periods, [7] 100 cycles of 1,000 cps modulation pulses, [8] In relation to WWV, [9] Interruption from the 15th to the 20th minutes of each hour, [10] Transmission on 60 ks also, [11] The 1st of the month, if necessary, [12] See carrier frequencies, [13] From 0700 to 2300 U.T., [14] Mondays, [15] Wednesdays, [16] Transmissions on 4 and 8 mc too, [17] Interruptions during 20 ms, [18] Maximum radiation: North-East and South-West, [19] Tuesdays, [20] From 0800 to 1100 and from 1300 to 1600 U.T., [21] 440 and 100 cps alternately, [22] Transmission by the Belgian Royal Observatory.

present time it seems safe to assume that the spectrum lines discussed above will be the first for which practical application will be found as frequency standards.

Standard Frequency Broadcasts

Standardized radio frequencies are now broadcast by a number of agencies in various nations,[57,58] and usually include time signals. Table I, above, provided by the International Radio Consultative Committee, through the courtesy of B. Decaux, International Radio Consultative Committee Study Group VII, gives the principal characteristics of standard-frequency and time-signal stations. This table is correct as of August, 1954.

Changes in WWV Transmissions[59]

The presently used method of adjustment of the frequency of WWV is a slight modification of the method described in a previous reference,[25] namely, that the frequency of the standard-frequency oscillator is steered to keep Universal Time as determined by the U. S. Naval Observatory, which advises WWV on regulation of the oscillator. The slight modification is that the frequency is readjusted by no more than 1×10^{-9} parts per day. The frequency of WWV is measured by the National Bureau of Standards at Boulder, Colorado, and the correction data are supplied for the adjustment of the transmitter. Tables of corrections to the broadcast time signals are furnished, as previously, by the Time Service, U. S. Naval Observatory.

The transmitters at WWV are using single-sideband transmission of tone modulation on some of the carrier frequencies. The carrier is radiated continuously by one transmitter unit, the sideband giving the tone modulation being generated from the same frequency-standard oscillator by appropriate frequency dividers, modulators, and filters, and then radiated through a separate antenna.

Precision Frequency Measuring Equipment

Extension of the frequency range and accuracy of precision frequency measuring equipment has, of necessity, been carried out to keep pace with the microwave measurement field and the improved stable oscillators described above.

[58] H. B. Law, "Standard frequency transmission equipment at Rugby radio station," *Proc. IEE*, vol. 102, part 3, pp. 166–173; March, 1955.

[59] U. S. Bureau of Standards Letter Circular LC 1009, and Supplement; December 1, 1954.

Precision Standard-Frequency Calibrators

As was stated in the section of this paper devoted to time standards, the exact calibration of a quartz-crystal-controlled clock in terms of time is the only method now available for establishing an accurate frequency calibration of the oscillator driving the clock. The accuracy of a frequency measurement carried out by comparison with astronomical time measurements has been limited in the past by the errors in the measurements of time, by the fluctuations in the rate of rotation of the earth itself, and by the fluctuations in the rate of the clock driven by the crystal-controlled oscillator.[5,7,17,25] Clock stability having now been improved by a significant amount, it is expected that the new methods of astronomical observation (see Dual-Rate Moon Position Camera, above) and improvements on the standard methods of observation (improved photographic zenith tube) will result in better data on the relative variations of the variable factors.

In order to provide the high-stability clocks described above, it has been found essential to maintain several quartz-crystal clocks in a frequency-standard installation, and to intercompare these clocks to establish their performance as to relative rate and acceleration, i.e., their rates relative to each other. Current practice for such intercomparison in the United States appears to favor the use of one frequency-standard oscillator slightly off-set from the correct standard frequency to produce beat-notes with the other correctly adjusted, standard-frequency oscillators. Such a system then permits measuring and recording of the relative frequencies of the various oscillators by measuring and recording the beat-note frequency. The precision of measurement of such a system may then be increased by multiplying the frequencies of the oscillators to be compared, and using the beat-note measuring equipment as before.[9,25,60] Beat frequency measuring equipment has been constructed using digital electronic counters to measure the duration of a beat cycle between two standard oscillators, and to record this duration as a voltage produced by a suitable resistance-bridge circuit.[61,62]

Other methods of measurement involving comparison of frequencies have been devised. One system makes use of a frequency-multiplier stage multiplying the frequency, f_1, of the oscillator to be measured, by 10, and of a similar multiplier stage for multiplying the frequency of the reference standard, f_2, by 9. The two signals, $10f_1$ and $9f_2$, are then beat together, the beat-note being at approximately the frequency of f_1 or f_2 but containing 10 times the error of f_1 and 9 times the error of f_2. This process is then repeated except that the original $9f_2$ signal is used to heterodyne the 10th harmonic of the first beat note. By continuing this process on to the desired point, and subtracting out the original f_2 frequency in the final beating process, the error frequency can be multiplied sufficiently to increase the sensitivity of indication of the frequency change to the required degree. Recording may then be accomplished by utilizing commercially-available recording-type frequency meters.[26]

An interesting variation on these methods makes use of an off-set reference frequency produced by means of a rotary phase-shifter capable of continuous rotation. This phase-shifter is driven at a constant rate by a synchronous-motor-drive operated by the frequency standard, the input frequency from the reference standard thus being shifted by 1 cycle per second for each revolution-per-second of the 360 degree phase shifter. The unknown frequency is then heterodyned by this shifted standard frequency, which has been multiplied to the appropriate value, and the resulting beat note recorded as above.[62]

Although the methods of frequency measurement described above are those most recently described, spark chronographs and other electric time recorders are still widely used, and integrating phase meters, similar to the polyphase modulator device described by Marrison,[9] are sometimes used for comparing the relative frequencies of frequency standard oscillators.

Microwave Frequency Measuring Equipment

Accurate measurements of frequencies in the microwave range require apparatus for the generation of standard frequencies and for comparison of these frequencies with the unknown frequencies to be measured, with appropriate interpolating equipment to provide accurate measurement over a continuous range of frequencies. Apparatus for precision frequency measurement in the microwave region generally includes (1) frequency multipliers or harmonic generators to produce harmonics of known standard frequencies, and (2) a receiver or detector for mixing the unknown signal with the standard frequency in order to produce a beat frequency, which is then measured by (3) an interpolation system.[18,63–65] Application of frequency-scanning or spectrum-analyzer techniques to the detector unit has been used to improve ease of operation. Digital electronic counters have been applied to the problem of measuring the beat-note for interpolation purposes.

The most effective way presently available for generating microwave harmonics of standard frequencies appears to be by means of the use of crystal diodes as harmonic generators.[61–66] The driving power for a crystal-diode harmonic generator is usually furnished by a conventional negative-grid vacuum-tube frequency-multiplier chain,[61] although klystrons are used at the extreme end of the range.[65] Application of crystal-diode

[60] J. M. Shaull, "High precision automatic frequency comparator and recorder," *Tele-Tech*, vol. 14, pp. 58 ff.; January, 1955.

[61] J. M. Shaull, "Frequency multipliers and converters for measurement and control," *Tele-Tech*, vol. 14, pp. 86 ff.; April, 1955.

[62] J. McA. Steele, "The standard frequency monitor at the national physical laboratory," *Proc. IEE*, vol. 102, part 3, pp. 155–165 (with discussion); March, 1955.

[63] R. G. Talpey and Harold Goldberg, "A microwave frequency standard," Proc. IRE, vol. 35, pp. 965–969; September, 1947.

[64] C. G. Montgomery, Ed., "Technique of Microwave Measurements," McGraw-Hill Book Co., New York, N. Y., pp. 343–375; 1947.

[65] L. J. Rueger and A. E. Wilson, "The microwave frequency standard," *Radio-Electronic Engrg*, pp. 5-ff.; March, 1953.

[66] F. D. Lewis, "Harmonic generation in the U-H-F region by means of germanium crystal diodes," *Gen. Rad. Experimenter*, vol.

harmonic generators has produced some relatively simple calibrating equipment covering frequencies up to 10,000 mc. (Model 100, Presto Recording Corp., Paramus, New Jersey).

The use of locked-oscillators in frequency-multiplier systems has been extended to the microwave range, one piece of apparatus of this type designed specifically for microwave measurement purposes now commercially available (Model FM-4, Gertsch Products Inc., Los Angeles, California).

Frequency Dividers

Although many frequency measurement systems require frequency multipliers to reach the microwave region, it is also possible to use a microwave oscillator as a source and to divide its frequency for the operation of auxiliary measuring equipment, such as interpolation systems, and clock mechanisms. The regenerative-modulator divider circuit[67] appears to be well suited to use with presently available microwave components.[68] Frequency divider systems operating at lower frequencies can have a wider choice of circuits, regenerative-modulator dividers,[9] multivibrators,[20] and counter-type dividers[69–71] being widely used.

Decade Frequency Generators

Standard-frequency oscillators of extremely high stability are usually constructed in such a manner that their frequency of operation can be adjusted by relatively small amounts only.[36] Hence for measurement purposes, it is desirable to be able to generate frequencies controlled by the reference standard oscillator in order to provide known standard frequencies in the region in which it is desired to make measurements.

The easiest solution to this problem requires only a harmonic generator, or distorter, which can be tuned to the harmonic desired. This solution is usually inadequate for general measurement purposes since only a narrow range is covered at any one harmonic, and the exact calibration of this range must be established during the measurement. Furthermore, even though the range covered is narrow, the harmonics of lower-frequency stages of the system frequently interfere to cause ambiguity and difficulty in identification of the harmonic actually desired.

If the entire range of harmonics of a standard frequency is available simultaneously, it is usually possible to count the intervals from a known reference point. This system is widely used in commercial frequency-standard apparatus.

As the maximum frequency range of measurements has increased, techniques for improving the facility of identification of a given harmonic frequency have been developed. These techniques have taken the form of tuned selective circuits of narrow bandwidth for selecting an individual harmonic,[72] and of relatively complex systems of harmonic generation, harmonic selection, mixing, and filtering to generate a given frequency relatively free from spurious components. Commercial models of this type of standardized-decade-frequency generator[73] have been produced having good rejection of spurious beat notes and unwanted modulation components.

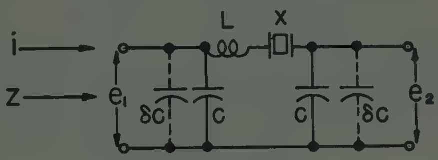


Fig. 19—Case I.

Appendix

Case I (Fig. 19)

C, C = Shunt capacitive elements (assumed equal).

L = Series inductance to bring crystal to series resonance when e_2 is 180 degrees out of phase with i.

X = Crystal.

δC, δC = Output and input capacitances of driving and driven tubes (assumed equal).

i = Input current.

e_1 = Input voltage developed.

e_2 = Output voltage.

Crystal Parameters.[74]

$Q_x = 2.6 \times 10^6$,

$R_x = 100\ \Omega$,

$L_x = 8.27\ h$,

$C_x = 0.000122\ \mu\mu f$,

$f = 5$ mc.

Circuit Analysis (Assuming crystal operating at series resonance). Let:

$$X_L = \omega L = 2X_c$$

where X_L is reactance of L, X_2 is reactance of one capacitance, C. The resistance of L is assumed small enough to be neglected.

$$B = \omega(C + \delta C).$$

Then

$$e_1 = \frac{i}{iB} \frac{(1 - BX_L) + jBR_x}{(2 - BX_L) + jBR_x}$$

$$e_2 = \frac{i}{B} \frac{1}{-BR_x + j(2 - BX_L)}.$$

[67] R. L. Miller, "Fractional frequency generation utilizing regenerative modulation," Proc. IRE, vol. 27, pp. 446–457; July, 1939.

[68] H. Lyons, "Microwave frequency dividers," *Jour. Appl. Phys.*, vol. 21, pp. 59–60; January, 1950.

[69] R. W. Frank, "A computer-type decade frequency synthesizer," 1954 IRE Convention Record, Part 10, "Instrumentation and Industrial Electronics," p. 46; 1954.

[70] R. W. Stuart, "A high speed digital frequency divider of arbitrary scale," 1954 IRE Convention Record, Part 10, "Instrumentation and Industrial Electronics," p. 52; 1954.

[71] G. K. Jensen and J. E. McGeogh, "Four-decade frequency divider," *Electronics*, vol. 28, pp. 154–155; April, 1955.

[72] J. M. Shaull, "Wide range decade frequency generator," *Tele-Tech*, vol. 9, p. 36; November, 1950.

[73] The Plessey Co., Ltd., Ilford (Essex), Eng.; A. Schomandl, Munich, Germany; Rohde and Schwarz, Munich, Germany; Telefunken, A. G., Berlin, Germany.

[74] A. W. Warner, "High-frequency crystal units for primary frequency standards," Proc. IRE, vol. 40, pp. 1030–1033; Sept. 1952.

For e_2 180 degrees out of phase with i

$$B = \frac{2}{X_L}$$

$$\frac{e_1}{i} = Z = \frac{X_L^2}{4R_x} - j\frac{X_L}{2}$$

$$\frac{e_2}{i} = -\frac{X_L^2}{4R_x}.$$

Numerical Values.
Let

$$\frac{e_2}{i} = -\frac{X_L^2}{4R_x} = -\frac{100}{3}.$$

(This value of transfer impedance is also satisfactory for Case II, thus enabling direct comparison.) Then:

$$X_L = 115.5\ \Omega$$

$$L = 3.68\ \mu\text{h}$$

If the Q of L is 230, which is reasonable for a coil of this inductance at this frequency, then

$$X_L/R_L = 230 = 115.5/R_L$$

$$R_L = \frac{115.5}{230} \approx 0.5\ \Omega,$$

which is negligible, as assumed above.

$$C + \delta C = 552\ \mu\mu\text{f}.$$

Assume 0.1 per cent change in L:

$$X_L = 115.5 \times 10^{-3} = 0.1155\ \Omega.$$

This change in reactance must be balanced by a change in crystal reactance to correct phase back to original value. This requires a small change of frequency, Δf. For small changes of frequency close to the series resonance frequency, the crystal reactance

$$X_q = X_0\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)$$

where X_0 is reactance of crystal inductance, X_L, at series-resonance frequency; $X_0 = 2(5\times10^6)\times8.27 = 2.6 \times 10^6\Omega$:

$$X_q = X_0\left(\frac{f+\Delta f}{f} - \frac{f}{f+\Delta f}\right)$$

$$X_0\frac{(f+\Delta f)^2 - f^2}{f(f+\Delta f)} \cong X_0\frac{2f\Delta f + \Delta f^2}{f(f+\Delta f)}.$$

Neglecting higher order terms,

$$X_q = X_0\frac{2\Delta f}{f}$$

$$\frac{X_q}{2X_0} = \frac{\Delta f}{f}$$

X_q then must equal ΔX_L:

$$\frac{\Delta f}{f} = \frac{\Delta X_L}{2X_0} = \frac{0.1155}{2.6 \times 10^6} = 2 \times 10^{-10}.$$

Assume 0.1 per cent change in each shunt capacitance, C:

$$\frac{\Delta f}{f} = 2 \times 10^{-10}.$$

(This is equivalent to 0.5 $\mu\mu$f in each tube capacitance.) Assume 1 $\mu\mu f$ change in one tube capacitance:

$$\frac{\Delta f}{f} = 2 \times 10^{-10}.$$

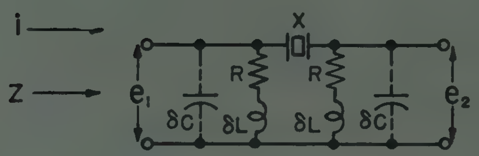


Fig. 20—Case II.

Case II (Fig. 20)

R, R=Shunt resistive elements, X=crystal, δC, δC=output and input capacitances of driving and driven tubes, assumed equal; δL, δL=compensating inductances, i=input current, e_1=input voltage developed, and e_2=output voltage.

Crystal Parameters. Same as Case I.

Circuit Analysis. e_2 will be in phase with i when crystal is at series resonance if the shunt impedances are both resistive. This occurs when

$$L = R^2\delta C$$

$$\frac{e_1}{i} = \frac{R(R+R_x)}{2R+R_x}$$

$$\frac{e_2}{i} = \frac{R^2}{2R+R_x}.$$

Numerical Values. Let:

$$\frac{e_2}{i} = \frac{R^2}{2R+R_x} = \frac{100}{3}.$$

Then:

$$R = 100\ \Omega.$$

Assume:

$$C = 10\ \mu\mu\text{f},$$

then

$$L = 0.1\ \mu\text{h} = R^2C.$$

Assume 0.1 per cent change in each shunt resistance, R:

$$\frac{\Delta(2R^2\delta C)}{L_x} = 4.8 \times 10^{-11},$$

$$\frac{\Delta f}{f} = 2 \times 10^{-11}.$$

Assume 0.1 per cent change in each compensating inductance, δL:

$$\frac{\Delta f}{f} = 1 \times 10^{-11}.$$

Assume 1 $\mu\mu$f change in one tube capacitance, δC:

$$\frac{\Delta f}{f} = 6 \times 10^{-10}.$$

Recapitulation

On the assumption that these circuits operate so that the crystal is at series resonance, and that the transfer impedance e_2/i is the same for both (100/3), there is little to choose between them. Case II is less sensitive to changes in circuit constants than Case I by an order of magnitude (2×10^{-11} vs 2×10^{-10}), but Case I is less sensitive to changes in tube capacitance by half an order of magnitude (2×10^{-10} vs 6×10^{-10}).

If over-all stability of 10^{-9} is assumed to be about all that can be reasonably expected, the frequency variations ascribed to the crystal coupling network and associated tube capacitances therefore do not seem to present a problem in either circuit. The next part of the analysis is devoted to the remaining part of the closed loop.

Loop Closure

Assume: Transconductance of tubes $= g_m = 1{,}000$ μmho $= 10^{-3}$. Then:

Gain of crystal-coupling-circuit portion

$$= \frac{100}{3} \times 10^{-3} = \frac{1}{30}$$

from grid of driving tube to grid of driven tube. Gain of remainder of closed loop must therefore $= 30$.

Assume: Two tubes, coupled through simple parallel-resonant circuit, transconductance of tubes $= g_m = 1{,}000$ μmho $= 10^{-3}$.

$$\text{Gain} = 30 = R_\beta g_m = 10^{-3} R_\beta$$

$$R_\beta = 30\ k\Omega = 3 \times 10^4$$

where R_β = impedance of coupling network at resonance.

Assume: Interstage capacitance $= 20$ $\mu\mu$f, coil resonant with this capacitance.

$$C_\beta = 20\ \mu\mu\text{f}$$

$$X_\beta = 1{,}592\ \Omega \text{ at } 5\text{ mc}$$

$$L_\beta = 50.7\ \mu\text{h}$$

$$Q_\beta = 18.9 \text{ at } 5\text{ mc}$$

$$\tan\theta_\beta = Q_\beta\left(\frac{\omega_\beta}{\omega} - \frac{\omega}{\omega_\beta}\right),$$

where R_β, C_β and L_β are tuned-circuit parameters, Q_β is the storage factor at the resonant frequency $f_\beta = \omega_\beta/2\pi$, and θ_β is the phase angle of the coupling system.

Assume: 1 $\mu\mu$f change in one tube capacitance

$$\frac{\Delta\omega_\beta}{\omega_\beta} = 2.5 \times 10^{-2}$$

$\tan\theta_\beta = 0.943$ (actual frequency $f = \omega/2\pi$ assumed constant). For crystal coupling network,

$$\tan\theta_\mu = Q_\mu\left(\frac{\omega_\mu}{\omega} - \frac{\omega}{\omega_\mu}\right),$$

where Q_μ is the storage factor at the resonant frequency $f_\mu = \omega_\mu/2\pi$, and θ_μ is the effective phase angle of the crystal coupling network.

Case I

$$Q_\mu = Q_x = 2.6 \times 10^6$$

$$\frac{\Delta f}{f} = 2 \times 10^{-7}$$

Case II

$$Q_\mu = \frac{2.6 \times 10^6}{3}$$

$$\frac{\Delta f}{f} = 6 \times 10^{-7}.$$

Recapitulation. Frequency shift from change in phase-shift of the loop is more important than changes in the crystal coupling network by three orders of magnitude (6×10^{-7} vs 6×10^{-10}). Case I is less sensitive than Case II to changes in capacitance in the closing loop by a half order of magnitude because a factor of three in effective Q_μ is sacrificed in Case II to work the crystal in and out of shunt resistive elements. Both circuits, however, are seriously limited by phase-shift in the closing loop.

It was noted that the coupling circuit was tuned entirely by the interstage capacitance, but this assumption need not be made. If additional capacitance is added at this point, the effect of a change in tube capacitance on the resonant frequency will be reduced, but the storage factor, Q_β, and consequently the rate of change of phase with frequency will be increased to the same extent. The phase shift introduced by a given change in tube capacitance will therefore remain the same, whether or not additional shunt capacitance is employed. If no extra capacitance is added the effect of any change in inductance is a minimum, however, and can be ignored.

It should be noted that in Case II there is zero phase shift in the crystal coupling network, whereas in Case I there is 180 degrees phase shift. Case II is therefore more readily adaptable to two-tube operation. A reasonably simple solution for Case I might be the use of a cathode-coupled twin triode for one of the two tubes.

Coupling systems designed for lower rate of change of phase shift might be worked out, but it would seem a

more promising avenue of approach to eliminate the network entirely by going to a single-tube circuit in which the driven and driving tube for the crystal coupling network were one and the same.

Single-Tube Version

To make a single-tube version, the reverse problem exists regarding phase-shift in the Case II and Case I circuits. Case I is more readily adaptable than Case II because of its 180 degree phase shift in the crystal coupling network. To make the Case II circuit work it would be necessary to go to some such expedient as use of a cathode-coupled twin triode.

Since there is no additional gain provided elsewhere, the gain from the grid of the "driving" tube to the grid of the "driven" tube must be unity (actually the same grid), and this specification therefore determines the transfer impedance of the crystal coupling network in terms of the tube transconductance. Assume:

$$\text{Transconductance} = g_m = 1{,}000\ \mu\text{mho} = 10^{-3}$$
$$i = i_p = -g_m e_g = -10^{-3} e_g = -10^{-3} e_2.$$

Case III

$$\frac{e_2}{i} = -10^3 = \frac{-X_L^2}{4R_x},$$
$$X_L = 632\ \Omega$$
$$L = 20.1\ \mu\text{h}$$
$$C, C = 101\ \mu\mu\text{f}$$

Assume 0.1 per cent change in L:

$$\frac{\Delta f}{f} = 1 \times 10^{-9}.$$

Assume 0.1 per cent change in each shunt capacitance, C:

$$\frac{\Delta f}{f} = 1 \times 10^{-9}.$$

Assume 1 $\mu\mu$f change in one tube capacitance:

$$\frac{\Delta f}{f} = 6 \times 10^{-9}.$$

Case IV (Two-Tube Circuit)

$$\frac{e_2}{i} = 10^3 = \frac{R^2}{2R + R_x}$$
$$R = 2{,}050\ \Omega$$
$$\delta L = 42\ \mu\text{h} = R^2 \delta C.$$

Assume 0.1 per cent change in each shunt resistance, R:

$$\frac{\Delta(2R^2\delta C)}{L_x} = 2 \times 10^{-8}$$
$$\frac{\Delta f}{f} = 1 \times 10^{-8}.$$

Assume 0.1 per cent change in each compensating inductance, δL:

$$\frac{\Delta f}{f} = 5 \times 10^{-9}.$$

Assume 1 $\mu\mu$f change in one tube capacitance, δC:

$$\frac{\Delta f}{f} = 3 \times 10^{-7}.$$

Recapitulation. In the single-tube version, Case I is markedly superior to the Case II circuit. To obtain the necessary gain, the impedance level of the crystal coupling network of Case II becomes too high, and dependence of frequency upon circuit parameters is substantial. The worst variation comes from changes arising from variations in tube capacitance, which are worse than those in Case I by almost two orders of magnitude.

In the single-tube, Case I, oscillator (Case III), however, the changes in frequency from this source are still only 6×10^{-9}, and it seems probable that sensible circuit design could reduce tube capacitance variations to about 0.1 $\mu\mu$f, rather than 1 $\mu\mu$f. Requirements of 0.1 per cent stability in circuit parameters are not unreasonable, and it therefore seems feasible to construct an oscillator of this type to yield circuit stability of 10^{-9}.

The advantage of using dc control of effective g_m for amplitude control, rather than a thermal bridge, is indicated by the sensitivity of the frequency to phase shift in circuits other than the crystal coupling network. Anything that reduces gain around the loop requires increased gain elsewhere, and this gain can only be obtained at the expense of great care in maintaining low rate of change of phase shift. It seems probable that the simple Case III circuit, using one oscillator tube and a stable, amplified, delayed AVC system with semi-starved operation of the oscillator tube will give not only an inexpensive solution but, perhaps, the best one.

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Applicator (Applicator Electrodes), (*Dielectric Heating* usage). Appropriately shaped conducting surfaces between which is established an alternating electric field for the purpose of producing *dielectric heating*.

Applicator Impedance, Loaded (*Dielectric Heating* usage). See *Loaded Applicator Impedance*.

Applicator Impedance, Unloaded (*Dielectric Heating* usage). See *Unloaded Applicator Impedance*.

Autoregulation Induction Heater. An induction heater in which a desired control is effected by the change in characteristics of a magnetic charge as it is heated at or near its *Curie point*.

Channel, Melting. See *Melting Channel*.

Charge. See *Load* (*Induction* and *Dielectric Heating* usage).

Contactor, Load. See *Load Switch* (*Load Contactor*).

Converter, Mercury Arc, Pool Cathode. See *Pool Cathode Mercury Arc Converter*.

Converter, Quenched Spark Gap. See *Quenched Spark Gap Converter*.

Converter, Mercury Hydrogen Spark Gap. See *Mercury Hydrogen Spark Gap Converter*.

Core Type Induction Heater or Furnace. A device in which a *charge* is heated by induction and a magnetic core links the inducing winding with the charge.

Coreless Type Induction Heater or Furnace. A device in which a charge is heated by induction and no magnetic core material links the charge.

Note—Magnetic material may be used elsewhere in the assembly for *flux guiding* purposes.

Coupling (*Induction Heating* usage). The percentage of the total magnetic flux produced by an inductor which is effective in heating a *load* or *charge*.

Curie Point (*Induction Heating* usage). The temperature in a ferromagnetic material above which the material becomes substantially nonmagnetic.

Decalescent Point (of a metal). The temperature at which there is a sudden absorption of heat as the metal is raised in temperature.

Depth of Heating (*Dielectric Heating* usage). The depth below the surface of a material in which effective *dielectric heating* can be confined when the *applicator electrodes* are applied adjacent to one surface only.

Depth of Penetration (*Induction Heating* usage). The thickness of a layer extending inward from the surface of a conductor, which has the same resistance to direct current as the conductor as a whole has to alternating current of a given frequency.

Note—This term is useful only in cases where the surface is substantially flat.

Dielectric Dissipation Factor. The cotangent of the *dielectric phase angle* of a dielectric material.

Dielectric Heating. The heating of a nominally insulating material in an alternating electric field due to its internal losses.

Dielectric Phase Angle. The angular difference in phase between the sinusoidal alternating voltage applied to a dielectric and the component of the resulting alternating current having the same period as the voltage.

Dielectric Power Factor. The cosine of the *dielectric phase angle*.

Dielectric Strength. The maximum potential gradient that a material can withstand without rupture.

Domestic Induction Heater. A cooking device in which the utensil is heated by current, usually of commercial line frequency, induced in it by a primary inductor associated with it.

Dual Frequency Induction Heater or Furnace. A heater in which the *charge* receives energy by induction, simultaneously or successively, from a *work coil* or coils operating at two different frequencies.

Efficiency, Over-all Electrical. See *Over-all Electrical Efficiency* (*Induction* and *Dielectric Heating* usage).

Efficiency, Load Circuit. See *Load Circuit Efficiency* (*Induction* and *Dielectric Heating* usage).

Field Strength Meter. A calibrated radio receiver for measuring field strength.

Flux Guide (*Induction Heating* usage). Magnetic material to guide electromagnetic flux in desired pahts.

Note—The guides may be used either to direct flux to preferred locations or to prevent the flux from spreading beyond definite regions.

Gaseous Tube Generator. A power source comprising a gas-filled electron tube oscillator, a power supply, and associated control equipment.

Glue Line Heating (*Dielectric Heating* usage). An arrangement of electrodes designed to give preferential heating to a thin film of material of relatively high loss factor between alternate layers of relatively low loss factor.

Heater Coil. See *Load Coil* (*Induction Heating* usage).

Heating Pattern. The distribution of temperature in a *load* or *charge*.

Heating Station. Location which includes *work coil* or *applicator* and its associated production equipment.

High-Frequency Induction Heater or Furnace. A device for causing electric current flow in a *charge* to be heated, the frequency of the current being higher than that customarily distributed over commercial networks.

Horizontal Ring Induction Furnace. A device for melting metal comprising an angular horizontally-placed open trough or *melting channel*, a primary inductor winding and a magnetic core which links the *melting channel* with the primary winding.

Hysteresis Heater. An induction device in which a *charge* or a muffle about the *charge* is heated principally by hysteresis losses due to a magnetic flux which is produced in it.

Note—A distinction should be made between *hysteresis heating* and the enhanced *induction heating* in a *magnetic charge.*

Induced Current (*Induction Heating* usage). Current in a conductor due to the application of a time-varying electro magnetic field.

Induction-Conduction Heater. A heating device in which electric current is conducted through but is restricted by induction to a preferred path in a charge.

Induction Heating. The heating of a nominally conducting material in a varying electro magnetic field due to its internal losses.

Induction Ring Heater. A form of *core-type induction heater* adapted principally for heating electrically conducting *charges* of ring or loop form, the core being open or separable to facilitate linking the *charge.*

Interference (*Induction* or *Dielectric Heating* usage). The disturbance of any electric circuit carrying intelligence, caused by the transfer of energy from an *induction* or *dielectric heating* equipment.

Load (*Induction* and *Dielectric Heating* usage) (*Charge*). The material to be heated.

Load Circuit (*Induction* and *Dielectric Heating* usage). The network including leads connected to the output terminals of the generator.

Note—The load circuit consists of the coupling network and the load material at the proper position for heating.

Load Circuit Efficiency (*Induction* and *Dielectric Heating* usage). The ratio of the power absorbed by the *load* to the power delivered at the generator output terminals.

Load Coil (*Induction Heating* usage). An electric conductor which, when energized with alternating current, is adapted to deliver energy by induction to a charge to be heated.

Load Leads (*Induction* and *Dielectric Heating* usage). The connections or transmission line between the power source or generator and *load*, *load coil* or *applicator.*

Load Matching (*Induction* and *Dielectric Heating* usage). The process of adjustment of the *load circuit* impedance to produce the desired energy transfer from the power source to the *load.*

Load Matching Network (*Induction* and *Dielectric Heating* usage). An electric network for accomplishing *load matching.*

Load Matching Switch (*Induction* and *Dielectric Heating* usage). A switch in the *load matching network* to alter its characteristics to compensate for some sudden change in the *load* characteristics, such as passing through the *Curie point.*

Load Switch (*Load Contactor*). The switch or contactor in an *induction heating* circuit which connects the high-frequency generator or power source to the *heater coil* or *load circuit.*

Load Transfer Switch. A switch to connect a generator or power source optionally to one or another *load circuit.*

Loaded Applicator Impedance (*Dielectric Heating* usage). The complex impedance measured at the point of application with the load material at the proper position for heating, at a specified frequency.

Low-Frequency Induction Heater or **Furnace.** A device for inducing current flow of commercial power line frequency in a *charge* to be heated.

Magnetron. An electron tube characterized by the interaction of electrons with the electric field of a circuit element in crossed steady electric and magnetic fields to produce ac power output.

Melting Channel. The restricted portion of the *charge* in a *submerged resistor* or *horizontal ring induction furnace* in which the induced currents are concentrated to effect high energy absorption and melting of the *charge.*

Mercury Arc Converter, Pool Cathode. See *Pool Cathode Mercury Arc Converter.*

Mercury Hydrogen Spark Gap Converter. A spark gap generator or power source which utilizes the oscillatory discharge of a capacitor through an inductor and a spark gap as a source of radio-frequency power. The spark gap comprises a solid electrode and a pool of mercury in a hydrogen atmosphere.

Motor Effect. The repulsion force exerted between adjacent conductors carrying currents in opposite directions.

Motor Field Induction Heater. An *induction heater* in which the inducing winding typifies that of an induction motor of rotary or linear design.

Oscillator. A nonrotating device for producing alternating current, the output frequency of which is determined by the characteristics of the device.

Over-all Electrical Efficiency (*Induction* and *Dielectric Heating* usage). The ratio of the power absorbed by the load material to the total power drawn from the supply lines.

Pad Electrode. One of a pair of electrode plates between which a *load* is placed for *dielectric heating*.

Pinch Effect. The result of an electromechanical force that constricts, and sometimes momentarily ruptures, a molten conductor carrying current at high density.

Pool Cathode Mercury Arc Converter. A frequency converter using a mercury arc power converter.

Proximity Effect. The redistribution of current in a conductor brought about by the presence of another conductor.

Quenched Spark Gap Converter. A spark gap generator or power source which utilizes the oscillatory discharge of a capacitor through an inductor and a spark gap as a source of radio frequency power. The spark gap comprises one or more closely-spaced gaps operating in series.

Radio Frequency Converter. A power source for producing electrical power at a frequency of 10 kc and above.

Radio Frequency Generator—Electron Tube Type (*Industrial* and *Dielectric Heating* usage). A power source comprising an electron tube oscillator, an amplifier if used, a power supply and associated control equipment.

Recalescent Point (of a metal). The temperature at which there is a sudden liberation of heat as the metal is lowered in temperature.

Rotary Generator (*Induction Heating* usage). An alternating-current generator adapted to be rotated by a motor or prime mover.

Shield. Material used to suppress the effect of an electric or magnetic field within or beyond definite regions.

Stirring Effect. The circulation in a molten *charge* due to the combined forces of *motor* and *pinch effects*.

Submerged Resistor Induction Furnace. A device for melting metal comprising a melting hearth, a depending *melting channel* closed through the hearth, a primary induction winding and a magnetic core which links the *melting channel* and the primary winding.

Unloaded Applicator Impedance (*Dielectric Heating* usage). The complex impedance measured at the point of application, without the *load* material in position, at a specified frequency.

Wave Heating. The heating of a material by energy absorption from a traveling electro magnetic wave.

Work Coil. See *Load Coil* (*Induction Heating* usage).

IRE Standards on Antennas and Waveguides: Definitions for Waveguide Components, 1955*

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Introduction

A set of definitions of basic waveguide terms prepared by the Technical Committee on Antennas and Waveguides was published as IRE Standards in December, 1953 (53 IRE 2. S1). The present Standards on Waveguide Components, Definitions of Terms, represent an extension of the work of this committee from 1953 to 1955.

Waveguide component terms have not previously been standardized by the IRE; however, with continuing development in this field the need for such definitions has become increasingly apparent. The present list comprises only the more general, basic and established terms. For example, *Waveguide Transformer* is defined but the many specific types of waveguide transformer such as *Double Stub Transformer*, *Quarter-Wave Sleeve Transformer*, *Probe Transformer*, *Eccentric Line Transformer*, and others are not defined.

As used in the following definitions, *Waveguide* is a generic term which includes *transmission line* and *uni-conductor waveguides* as special cases. For specific definitions see 53 IRE 2. S1.

Definitions

Attenuator, Waveguide. A waveguide device for the purpose of producing attenuation by any means, including absorption and reflection.

Bend, Waveguide. A section of waveguide in which the direction of the longitudinal axis is changed.

Butt Joint. A connection between two waveguides which provides physical contact between the ends of the waveguides in order to maintain electrical continuity.

Cavity Resonator (in Waveguides). A resonator formed by a volume of dielectric bounded by reflecting walls.

Cavity Resonator Frequency Meter. A cavity resonator used to determine frequency of an electromagnetic wave.

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Choke Joint. A connection between two waveguides which provides effective electrical continuity without metallic continuity at the inner walls of the waveguide.
Connector, Waveguide. A mechanical device for electrically joining separable parts of a waveguide system.
Coupling Aperture (Coupling Hole, Coupling Slot). Aperture in wall of waveguide or *cavity resonator* designed to transfer energy to or from an external circuit.
Coupling Loop. A conducting loop projecting into a waveguide or *cavity resonator*, designed to transfer energy to or from an external circuit.
Coupling Probe. A probe projecting into a waveguide or *cavity resonator* designed to transfer energy to or from an external circuit.
Directional Coupler. A four-branch junction consisting of two waveguides coupled together in a manner such that a single traveling wave in either guide will induce a single traveling wave in the other, direction of latter wave being determined by direction of the former.
E-H Tuner. An *E-H tee* used for impedance transformation having two arms terminated in adjustable plungers.
E-H Tee. A junction composed of a combination of *E* and *H-plane tee* junctions having a common point of intersection with the main guide.
E-plane Bend. For a rectangular uniconductor waveguide operating in the dominant mode, a bend in which the longitudinal axis of the guide remains in a plane parallel to the electric field vector throughout the bend.
E-plane Tee Junction. For a rectangular uniconductor waveguide, a tee junction of which the electric field vector of the dominant wave of each arm is parallel to the plane of the longitudinal axes of the guides.
H-plane Bend. For a rectangular uniconductor waveguide operating in the dominant mode, a bend in which the longitudinal axis of the guide remains in a plane parallel to the plane of the magnetic field vector throughout the bend.
H-plane Tee Junction. For a rectangular uniconductor waveguide, a tee junction of which the magnetic field vector of the dominant wave of each arm is parallel to the plane of the longitudinal axes of the guides.
Hybrid Junction. Waveguide arrangement with four branches which, when branches are properly terminated, has the property that energy can be transferred from any one branch into only two of remaining three.

Note—In common usage, this energy is equally divided between the two branches.

Hybrid Tee. A *hybrid junction* composed of an *E-H Tee* with internal matching elements, which is reflectionless for a wave propagating into the junction from any arm when the other three arms are match terminated.
Iris (Diaphragm). In a waveguide, a conducting plate or plates, of thickness small compared to a wavelength, occupying a part of the cross section of the waveguide.

Note—When only a single mode can be supported an *iris* acts substantially as a shunt admittance.

Line Stretcher. A section of waveguide whose physical length is variable.
Magic Tee. See *Hybrid Tee.*
Mode Filter. A selective device designed to pass energy along a waveguide in one or more modes of propagation and substantially reduce energy carried by other modes.
Mode Transducer (Mode Transformer). A device for transforming an electromagnetic wave from one mode of propagation to another.
Phase Shifter, Waveguide. A device for adjusting the phase of a particular field component (or current or voltage) at output of device relative to the phase of that field component (or current or voltage) at the input.
Plunger, Waveguide. In a waveguide, a longitudinally movable obstacle which reflects essentially all the incident energy.
Post, Waveguide. In a waveguide, a cylindrical rod placed in a transverse plane of the waveguide and behaving substantially as a shunt susceptance.
Resonator, Waveguide (Resonant Element). A waveguide device primarily intended for storing oscillating electromagnetic energy.
Rotating Joint. A coupling for transmission of electromagnetic energy between two waveguide structures designed to permit mechanical rotation of one structure.
Series Tee Junction. A *tee junction* having an equivalent circuit in which the impedance of the branch guide is predominantly in series with the impedance of the main guide at the junction.
Shunt Tee Junction. A *tee junction* having an equivalent circuit in which the impedance of the branch guide is predominantly in parallel with the impedance of the main guide at the junction.
Slug Tuner. A *waveguide tuner* containing one or more longitudinally adjustable pieces of metal or dielectric.
Stub, Waveguide. An auxiliary section of waveguide with an essentially nondissipative termination and joined at some angle with the main section of waveguide.
Taper, Waveguide. A section of *tapered waveguide.*
Tapered Waveguide. A waveguide in which a physical or electrical characteristic changes continuously with distance along the axis of the guide.
Tee Junction. A junction of waveguides in which the longitudinal guide axes form a T.

Note—The guide which continues through the junction is the main guide; the guide which terminates at a junction is the branch guide.

Transformer, Waveguide. A device, usually fixed, added to a waveguide for the purpose of impedance transformation.
Tuner, Waveguide. An adjustable device added to a waveguide for the purpose of impedance transformation.
Tuning Probe. An essentially lossless probe of adjustable penetration extending through the wall of the waveguide or *cavity resonator.*
Twist, Waveguide. A waveguide section in which there is a progressive rotation of the cross section about the longitudinal axis.
Wye Junction. A junction of waveguides such that the longitudinal guide axes form a Y.

High-Frequency Power Gain of Junction Transistors*

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Summary—The purpose of this paper is three-fold. First, the subject of maximum available power gain at high frequencies is discussed briefly. Also, maximum gain for a four-terminal network driven by a generator having a purely resistive internal impedance is calculated in terms of *small-signal parameters* of the network. Then a theoretical model of a junction transistor comprising the ideal one-dimensional model plus a base impedance, which may be complex and frequency-dependent as in the case of grown-junction transistors, is introduced for the network to obtain an expression for maximum available power gain in terms of fundamental *device parameters*. Experimental results, which are given for a number of grown-junction transistors, tend to confirm the theoretical expression. Finally, an idealized model of a grown-junction transistor is introduced, and theoretical power gain is calculated in terms of *physical parameters*. Such calculations show, for example, that 30 db of gain should be available at 5 mc and that such transistors should be capable of oscillating up to several hundred mc.

Introduction

THE SUBJECT of the high-frequency performance of a junction transistor has received considerable attention during the past few years. Variation of transistor parameters with frequency has been discussed in some detail,[1] and recently several writers have presented equations for relating high-frequency power gain to transistor parameters for transistors having constant base-spreading resistance.[2] The purpose of the present paper is three-fold. First, the different ways in which power gain at high frequencies may be defined is discussed briefly. Then an equation is presented for calculating high-frequency power gain in terms of *four-pole* parameters for a transistor which is conjugate-matched at the output and which is driven by a generator having a purely resistive internal impedance. Second, a new, different expression relating high-frequency gain to *fundamental device* parameters is presented for the case of junction transistors in which base-spreading resistance is *not constant* at high frequencies, e.g., as in *grown*-junction transistors. Results of measurements of power gain for approximately 60 grown-junction transistors are shown which tend to confirm the validity of this theoretical expression. Third, values of high-frequency power gain are calculated for an idealized theoretical model of a grown-junction transistor triode, in order to illustrate what upper limit exists on power gain from such transistors. For example, calculations show that 30 db of power gain should be available at 5 mc, and that such transistors should be capable of oscillating up to several hundred mc.

Definition of High-Frequency Power Gain

At high frequencies, there is no clear-cut interpretation for the maximum gain of a junction transistor, since under proper terminations a transistor may oscillate. Hence, maximum gain could be infinity. Accordingly, in general, gain must be maximized subject to certain constraints. One measure of power gain is the unilateral power gain, or U function, proposed by Mason.[3] On the other hand, J. G. Linvill has derived an expression for the maximum gain available from a general linear four-terminal network in terms of the series-parallel h parameters.[4] This expression yields a value of gain within 3 db of the maximum possible gain available, unless the transistor would oscillate under proper terminations. By an additional simple calculation, it is possible to determine whether or not oscillations could be obtained, i.e., whether or not the transistor is potentially unstable.

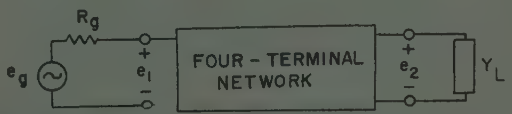

Fig. 1—Four-terminal network with variable load admittance and with generator having resistive internal impedance.

Alternatively, the writer has calculated the maximum gain available from a four-terminal network when driven by a generator having a purely *resistive* internal impedance, as shown in Fig. 1. The importance of such a calculation lies in the fact that experimental determination of high-frequency gain often is made as shown in Fig. 1.[5] By employing this type of measurement, it is

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[1] For example, J. M. Early, "Design theory of junction transistors," *Bell Sys. Tech. Jour.*, vol. 32, pp. 1271–1312; November, 1953.
R. L. Pritchard, "Frequency variations of junction-transistor parameters," Proc. IRE, vol. 42, pp. 786–799; May, 1954. Presented at Transistor Research Conference, State College, Pa.; July 6, 1953.
H. Johnson, RCA Laboratories, unpublished paper presented at Transistor Research Conference, State College, Pa.; July 6, 1953.

[2] R. L. Pritchard, "Frequency Response of Grounded-Base and Grounded-Emitter Junction Transistors," presented at AIEE Winter Meeting, New York; January 22, 1954.
J. M. Early, "*Pnip* and *npin* junction transistor triodes," *Bell Sys. Tech. Jour.*, vol. 33, p. 519; May, 1954.
L. J. Giacoletto, "The study and design of alloyed-junction transistors," 1954 IRE Convention Record, Part 3, p. 102. Also, "Study of *p-n-p* alloy junction-transistor from dc through medium frequencies," *RCA Rev.*, vol. 15, p. 555; December, 1954.
H. Statz, E. A. Guillemin, and R. A. Pucel, "Design considerations of junction transistors at higher frequencies," Proc. IRE, vol. 42, p. 1627; November, 1954.

[3] S. J. Mason, "Power gain in feedback amplifiers," Trans. IRE, vol. CT-1, pp. 20–25; June, 1954. Note, however, that Mason points out that under certain conditions (of lossy coupling) a transistor can yield more gain than that calculated from the U function (corresponding to lossless coupling).

[4] J. G. Linvill, "The Relationship of Transistor Parameters to Amplifier Performance," presented at IRE-AIEE—University of Pennsylvania Conference on Transistor Circuits, Philadelphia, Pa., February 17, 1955.

[5] See, for example, J. I. Pankove and C. W. Mueller, "A *p-n-p* triode alloy-junction transistor for radio-frequency amplification," Proc. IRE, vol. 42, p. 390; February, 1954; *RCA Rev.*, vol. 14, p. 596; December, 1953.

often possible to avoid oscillations which otherwise might result if an attempt is made to conjugate match at both output and input terminals simultaneously. The very important question of how much gain can be obtained with a potentially unstable transistor (gain vs stability question) is *not* considered in this paper. What *is* required here is a relationship between network parameters and network performance. For this purpose, (1), based on the circuit of Fig. 1, is quite satisfactory (although Linvill's result[4] is more general), and experimental results can be obtained easily with the circuit of Fig. 1. Furthermore, as noted below, under certain conditions (which are satisfied quite well by most junction transistors in grounded-emitter operation at *high* frequencies) the maximum gain available from the circuit of Fig. 1 will be within a few db of the maximum gain available from the transistor under *any* conditions of termination.

The maximum available power gain for the circuit of Fig. 1 is given by the equation (see Appendix A):

$$G_{av} = |h_{21}|^2/[2r_{11}g_{22}(1 + p_m) - H_r],\ p_m > 0, \quad (1)$$

where

$$r_{11} \equiv \mathrm{Re}\,(h_{11}), \quad g_{22} \equiv \mathrm{Re}\,(h_{22}), \quad H_r \equiv \mathrm{Re}\,(h_{12}h_{21}), \quad (2)$$

and

$$p_m \equiv \left\{1 + \left(\frac{x_{11}}{r_{11}}\right)^2 - \frac{H_r}{r_{11}g_{22}}\left[1 + \left(\frac{x_{11}}{r_{11}}\right)\left(\frac{H_i}{H_r}\right)\right]\right\}^{1/2}, \quad (3)$$

with

$$x_{11} \equiv \mathrm{Im}\,(h_{11}), \qquad H_i \equiv \mathrm{Im}\,(h_{12}h_{21}). \quad (4)$$

Here Re and Im denote real and imaginary part, respectively.

Physically, parameter p_m of (3) may be identified as the ratio of generator resistance R_{gm} required for maximum gain, with conjugate matching at output circuit, to short-circuit input resistance r_{11}; i.e., $p_m = (R_{gm}/r_{11})$.

It should be emphasized that (1) is valid *only* for $p_m > 0$. For the conditions that yield $p_m = 0$, the circuit determinant vanishes; this corresponds to the condition for oscillations in the circuit shown in Fig. 1. Infinite gain, i.e., oscillations, also may be obtained for other values of p_m, e.g., if the denominator of (1) vanishes. Consequently, since the parameter H_r in (3) above may be positive for a junction transistor,[6] under proper conditions a transistor may oscillate in the circuit shown in Fig. 1, even with a purely resistive generator impedance.

A few additional remarks concerning (1) and (3) may be in order. First, note that at low frequencies all reactive terms vanish, and (1), together with (3), reduces to the exact equation for calculating maximum gain of a purely resistive four-terminal network.[7] Second, note that at high frequencies, calculations for (3) may be simplified somewhat by neglecting terms involving the reactive part x_{11} of the short-circuit input impedance h_{11}. Physically, this would correspond to tuning out the *short-circuit* input reactance, i.e., to adding in series with R_g a reactance $-x_{11}$. Since the *actual* input reactance of the transistor may be quite different from x_{11}, this would *not* correspond to conjugate matching at the input. W. N. Coffey of this Laboratory has considered the resulting gain expression in some detail and has pointed out that[8] the value of gain obtained from (1) by setting $x_{11} = 0$ in (3) *also* represents the maximum gain available with a variable *complex* generator impedance Z_g and a variable load conductance G_L, with the open-circuit output susceptance b_{22} tuned out, i.e., $\mathrm{I_m}(Y_L) = -b_{22}$. Finally, it might be noted that if the short-circuit input reactance x_{11} is equal to zero or can be tuned out and if $(H_r/r_{11}g_{22})$ is small relative to unity, then p_m may be replaced by the first two terms of its series expansion, and (1) reduces to the expression given by Linvill.

High-Frequency Power Gain in Terms of Fundamental Transistor Parameters

In order to calculate available gain for a given transistor, it is necessary merely to substitute measured values of the four complex h parameters into the appropriate equation by which gain is to be defined. However, by considering an appropriate model of a junction transistor, it is possible to relate the h parameters to certain fundamental device parameters of the transistor and hence to obtain an expression for high-frequency gain in terms of these fundamental parameters. The model comprises the usual ideal one-dimensional transistor in series with a base *impedance* z_b' [Fig. 2(a), next page]. For most fused-junction transistors, z_b' is real and is the base spreading resistance r_b'. However, for grown-junction transistors, in general, the distributed nature of the transistor parameters in the transverse direction of the base region must be taken into account. Results of a theoretical analysis[9] of a two-dimensional (distributed) model have shown that under simplifying conditions, such a transistor may be represented by the model

[6] Generally, H_r is positive for grounded-emitter operation and is negative for grounded-base.

[7] Equations for calculating low-frequency maximum power gain have been given, in different form, by R. L. Wallace, Jr., and W. J. Pietenpol, "Some circuit properties and applications of *n-p-n* transistors," *Bell Sys. Tech. Jour.*, vol. 30, pp. 546–549; July, 1951; Proc. IRE, vol. 39, pp. 759–761; July, 1951. See also H. E. Hollmann, "Transistortheorie und Transistorschaltungen," *Arch. elekt. Übertragung*, vol. 7, p. 326; July, 1953.

[8] W. N. Coffey, Personal communications and unpublished memorandum.

[9] R. L. Pritchard and W. N. Coffey, "Small-signal parameters of grown-junction transistors at high frequencies," 1954 IRE Convention Record, Part 3, pp. 90–98. Also, R. L. Pritchard, "Theory of grown-junction transistor at high frequencies," presented at Semiconductor Device Research Conference, Minneapolis, Minn., June 29, 1954; planned for publication.

shown in Fig. 2(a), in which case z_b' may be complex and frequency dependent *at high frequencies*. Hence, the *two-dimensional* distributed model actually may be represented by the ideal *one-dimensional* model plus a complex base impedance.

The h parameters for the *ideal* transistor have been described previously in general terms in some detail.[1]

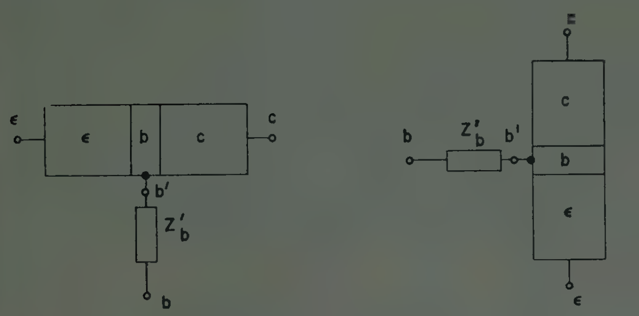

Fig. 2—Ideal transistor plus base impedance; (a) Grounded-base operation, (b) Grounded-emitter operation.

However, in order to obtain a relatively simple expression for maximum gain, certain assumptions can be made, consistent with what may be expected in commercially available transistors under practical operating conditions. For example, the emitter-efficiency term γ in the expression for current-amplification factor can be assumed to equal unity at all frequencies. Moreover, this assumption will be valid for practical transistors at not too high nor too low values of dc emitter current.[10]

Under such assumptions, fairly simple analytical expressions can be obtained for the h parameters for frequencies up to approximately twice the α-cutoff frequency. For example, see the equivalent circuit in Appendix B. Unfortunately, however, for grounded-base operation a relatively simple expression for gain does not result, although numerical calculations can be carried out fairly easily. Furthermore, if the subject of stability is investigated with the help of Linvill's gain-stability criterion,[4] it is found that the grounded-base configuration *may* be unstable at frequencies up to approximately the α-cutoff frequency.

On the other hand, if the grounded-emitter configuration is considered, a relatively simple, compact expression can be obtained for maximum gain. Furthermore, when Linvill's criterion is applied to this case, it is found that stable maximum gain can be obtained down to much lower frequencies.

For grounded-emitter operation, the h parameters of the transistor model which are shown in Fig. (2b) are as follows:

$$[h^{(e)}] = \begin{bmatrix} z_b' + h_{11}'^{(e)} & h_{12}'^{(e)} \\ h_{21}'^{(e)} & h_{22}'^{(e)} \end{bmatrix}, \tag{5}$$

where $h'^{(e)}$ denotes the grounded-emitter h parameter for the theoretical model. Note that the base impedance z_b' appears only in *one* parameter, viz., in h_{11}. General expressions for the grounded-emitter $h'^{(e)}$ parameters are given in Appendix B.

For calculating available power gain, the following simplifying assumptions will be made:

1. Emitter efficiency $\gamma = 1$ at all frequencies. This can be achieved in practice at not too high nor too low dc emitter currents.[10]
2. Limiting case of low frequencies is excluded, e.g., low-frequency interterminal conductances are neglected. This is equivalent to assuming $(1-\alpha_0)=0$. As a result of this assumption, the gain calculated according to the expression given below will be too high at low frequencies. Low-frequency gain can be calculated separately using completely different parameters, i.e., the low-frequency parameters, in order to determine when the high-frequency result becomes inapplicable.[7]
3. The effect of the base impedance is the dominant part of the input resistance r_{11}, i.e., $r_{11} \gg r_e'$, where $r_e' \equiv (kT/q_eI_e)$ is the Shockley, *et al.*, emitter resistance. This will be true at larger values of dc emitter current.
4. The collector-base diffusion capacity C_d is much smaller than the collector-base barrier capacity C_c. This will be true for not too large values of dc emitter current *density*. Hence, high-frequency output conductance g_{22} is predominantly due to the effect of collector-base barrier capacity C_c. Also, it follows that the feedback due to C_c is much larger than that due to Early effect at high frequencies, i.e., in the expression for $h_{12}^{(e)}$ (see Appendix B), $\mu_{ec} \ll 0.8\omega_c C_c r_e'$, where μ_{ec} is the Early feedback factor.[11]

Consideration of the assumptions shows that, in general, gain will be decreased at low emitter currents and at high emitter currents. Actually, assumptions 1, 3, and 4 may be removed at only a slight expense in complexity. This is done below for the case of the grown-junction transistor.

Under the above assumptions and over a limited frequency range,[12] with $(\omega/\omega_c) < 2$, where $\omega_c/2\pi$ is the α-

[10] At low emitter current, emitter-base barrier capacity causes a decrease in γ. On the other hand, at high emitter currents, γ may decrease because of high-level injection effects. High-level injection was discussed first by W. M. Webster, "On the variation of junction transistor current-amplification factor with emitter current," presented at Transistor Research Conference, State College, Pa., July 7, 1953; Proc. IRE, vol. 42, pp. 914–920; June, 1954. See also E. S. Rittner, "Extension of the theory of the junction transistor," *Phys. Rev.*, vol. 94, pp. 1161–1171; June 1, 1954.

[11] J. M. Early, "Effects of space-charge layer widening in junction transistors," Proc. IRE, vol. 40, p. 1403; November, 1952.

[12] Under the assumption that $(1-\alpha_0)=0$, (6) are valid down to zero frequency. However, in a practical transistor with $(1-\alpha_0)\neq 0$, (6) will be valid for $(1-\alpha_0) \ll (\omega/\omega_c) < 2$.

cutoff frequency, the following approximate expressions may be obtained for the parameters required in calculating gain (see Appendix B):

$$r_{11} \doteq \mathrm{Re}\,(z_b'), \qquad g_{22} \doteq 0.8\,\omega_c C_c,$$
$$|h_{21}|^2 \doteq (1.2\,\omega/\omega_c)^{-2},$$

and[13]

$$\mathrm{Re}\,(h_{12}h_{21}) \equiv H_r \doteq -\left[\frac{\mu_{ec}}{6} + (\omega_c C_c r_e')\frac{(\omega/\omega_c)^2}{15} + (\omega_c C_c r_e')^2\left(1+\frac{C_e}{3C_c}\right)\right], \tag{6}$$

where C_e is the emitter-base barrier capacity. As a consequence of assumptions 3 and 4 above,

$$-\frac{H_r}{r_{11}g_{22}} = \frac{\mu_{ec}}{5r_{11}\omega_c C_c} + \left(\frac{r_e'}{r_{11}}\right)\left[\frac{(\omega/\omega_c)^2}{12} + 1.2(\omega_c C_c r_e')\left(1+\frac{C_e}{3C_c}\right)\right] \ll 1. \tag{7}$$

For a transistor having a constant base spreading resistance r_b', e.g., a normal fused-junction transistor, $r_{11}=r_b'$, and $(x_{11}/r_{11})\ll 1$. Then (6) and (7) may be substituted in (1) to yield the now familiar expression,[2]

$$G_{av} = \frac{0.22}{\omega^2}\left(\frac{\omega_c}{r_b'C_c}\right) \approx \frac{1}{25f^2}\left(\frac{f_c}{r_b'C_c}\right)$$
$$0.05\text{–}0.1 < (\omega/\omega_c) < 2. \tag{8}$$

Grown-Junction Transistor

On the other hand, for the distributed model of the transistor, e.g., a *grown*-junction transistor with a base contact that approximates a line contact, analysis[9] indicates that under the assumptions described above,

$$r_{11} \doteq [R_b r_e'/2.4(\omega/\omega_c)]^{1/2}, \text{ provided}$$
$$\left[\frac{R_b(\omega/\omega_c)}{r_e'}\right]^{1/2} > 2, \tag{9}$$

where R_b is the transverse base resistance of the transistor. (For a tetrode transistor, R_b would be the base-base resistance.)

If this inequality is not satisfied, the grown-junction transistor behaves like a fused-junction transistor with a simple base-spreading resistance $r_b'=R_b/3$.[14] This will be true in general at low and medium frequencies. Accordingly it is convenient to substitute this *frequency-independent* value of r_b' for R_b (which is not generally known for a triode) in (9); this yields:

$$r_{11} \doteq [r_b' r_e'/0.8(\omega/\omega_c)]^{1/2} \qquad [r_b'(\omega/\omega_c)/r_e']^{1/2} > 1. \tag{9a}$$

In this case, (x_{11}/r_{11}) being $\doteq -1$, is not negligible, and the expression for p_m of (3) becomes a bit cumbersome. Hence, in order to obtain a simple expression for gain, it is convenient to assume that p_m *can* be approximated by unity in (3).[15] In this particular case, the resulting expression for gain becomes identical with that calculated from Linvill's expression (since $H_r/r_{11}g_{22}$ *is* negligible[13]).

Substitution of (9a) together with (6) in (1) and (2), with $p_m=1$, yields,

$$G_{av} \doteq \frac{0.2}{\omega^{3/2}}\,\frac{\omega_c^{1/2}}{C_c(r_b'r_e')^{1/2}}, \quad 0.05\text{–}0.1 < (\omega/\omega_c) < 2.$$
$$[r_b'(\omega/\omega_c)/r_e']^{1/2} > 1. \tag{10}$$

Comparison of this result (applicable for *grown*-junction transistors) with (8) (applicable for *fused*-junction transistors) shows that the available power gain at high frequencies still is dependent upon the same parameters, viz., medium-frequency base-spreading resistance r_b', collector-base capacity C_c, and α-cutoff frequency $\omega_c/2\pi$. Note, however, that the dependence is different for the distributed model (grown-junction transistor), e.g., gain at a given frequency is proportional to the *square root* of the α-cutoff frequency, and gain varies with frequency at the rate of 15 db per decade rather than at 20 db per decade. Note also that for the distributed model, gain varies inversely as $r_e'^{1/2}$, i.e., gain is directly proportional to the square root of dc emitter current I_e. Although this can be confirmed in practice, variation of gain with I_e also can result from second-order effects, e.g., when one or more of assumptions 1, 3, or 4 above is not strictly valid, or when r_b' varies with I_e.[14]

Second-Order Effects

In view of the theoretical dependence of gain upon dc emitter current I_e for the distributed-model transistor, it might be of interest to remove several of the restrictions imposed above in order to calculate what might be called second-order correction terms. In particular, if the effect of emitter capacity is taken into account for low I_e, and

[13] This result is complete only for assumption 2 that $(1-\alpha_0)=0$. In a practical transistor, a *considerably* more important contribution to H_r, especially at lower frequencies, arises from the factor $(1-\alpha_0)$. The contribution in this case is

$$H_r \doteq \frac{(1-\alpha_0)}{(\omega/\omega_c)^2}\,[(\omega_c C_c r_e') + 0.6\mu_{ec}], \quad (\omega/\omega_c)^2 \gg 0.6(1-\alpha_0)^2.$$

It should be emphasized that *this* term, which is positive, may be sufficiently large in magnitude to satisfy the conditions required for infinite gain, i.e., oscillations, in (1) and (3). Since this term decreases with increasing frequency, such oscillations are most likely to occur at moderately low frequencies relative to f_c. Accordingly, it is desirable to limit the validity of the equations given below for gain from a theoretical model of a junction transistor to frequencies *greater* than perhaps 0.05 ω_c–0.1ω_c (although perfectly stable gain *may* be obtainable at lower frequencies). Linvill's stability criterion applied to the theoretical model, yields the result that stable gain can be obtained for $(\omega/\omega_c) > 0.4(r_e'/r_{11})$.

[14] See Pritchard and Coffey, reference 9, p. 93. Actually, this substitution is valid only for very small dc base current. When a substantial transverse base current exists, e.g., as at higher values of dc emitter current I_e, a grown-junction triode behaves like an internally biased tetrode transistor (*ibid.*, p. 95), and r_b' decreases with increasing I_e. However, even in this case, the value of $3r_b$ may be taken as a measure of the *effective* transverse base resistance.

[15] Actually, if the more complete expression (3) for p_m is considered for the theoretical model, as it should be when comparing experimental results for gain determined with a purely resistive generator impedance, the following results may be obtained:

If H_r and H_i terms are negligible, then $p_m \doteq \sqrt{2}$, and actual gain will be approximately 0.8 db *less* than that calculated by setting $p_m=1$. If the H terms are not negligible, oscillations may be obtained; see reference 13.

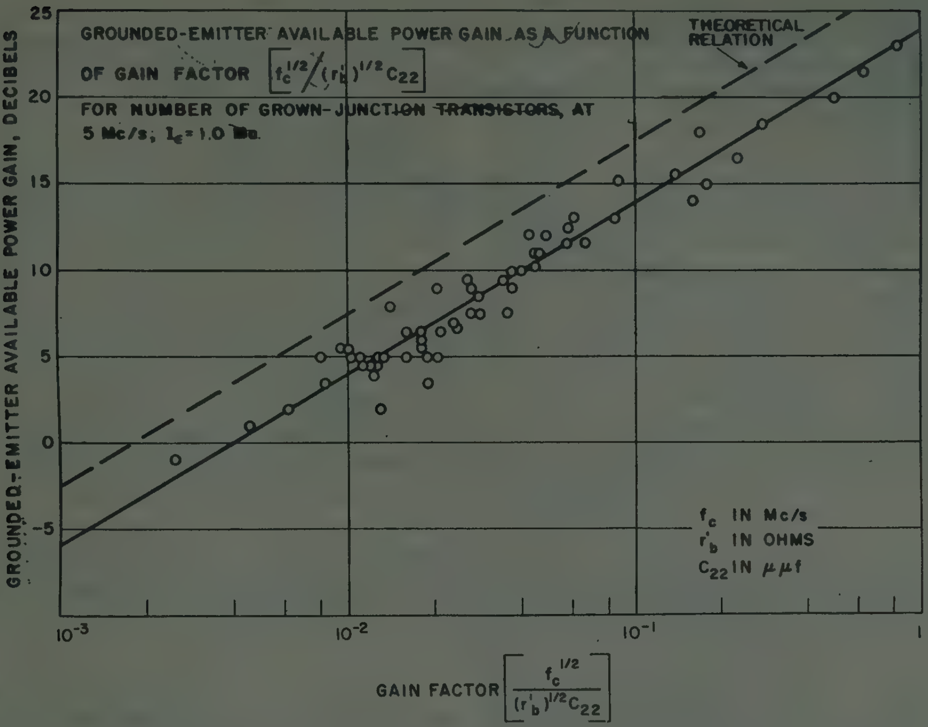

Fig. 3—Grounded-emitter maximum available power gain with resistive generator impedance as a function of gain factor $[f_c^{1/2}/(r_b')^{1/2}C_{22}]$; experimental results for a number of grown-junction transistors at 5 mc.

if the effect of collector-base diffusion admittance is taken into account at high emitter current *density*, the more complete expression for gain becomes

$$G \approx \frac{0.2}{\omega^{3/2}} \frac{\omega_c^{1/2}}{C_c(r_b'r_e')^{1/2}} \frac{1}{(1 + C_d/C_c)} \frac{1}{\sqrt{1+0.8\omega_c r_e' C_T}}. \quad (11)$$

In this equation, $C_T \equiv (C_e + C_c)$ is the sum of emitter-base and collector-base barrier capacities, and C_d is the (frequency-independent) collector-base diffusion capacity, due to the effect of space-charge-layer widening and the charge stored in the base region by the dc emitter current I_e. In particular, C_d is directly proportional to I_e and to Early's space-charge-layer widening factor, which in turn is a function of the nature of the collector junction.[16] Furthermore, it should be noted that the *ratio* of C_d/C_c is an indication of injection level.[17] High-level injection occurs when the injected minority-carrier charge density is comparable with the majority-carrier charge density normally present in the base. For $C_d/C_c \approx 1$, these two charge densities are approximately equal.

The expression given above is substantially complete except for the decrease of current gain at large values of dc emitter current, due to high-level injection effects. However, if it is assumed that the current gain is not adversely affected unless high-level injection is fairly substantial, then (11) *will* be valid up to approximately $C_d/C_c \approx 1$.[18] In this case, it might be noted that if the term involving C_T is negligible, a maximum power gain with respect to dc emitter current, with all other parameters held constant, occurs for $C_d/C_c = 1$.

Experimental Results

In an attempt to check the validity of (10), power gain and small-signal parameter measurements were made for a large number of grown-junction transistors originating from a number of different sources. Results for a dc emitter current $I_e = 1$ ma are shown in Fig. 3 in which each point corresponds to a different transistor. For each point, the ordinate value represents the maximum power gain available at a frequency of 5 mc from

[16] See R. L. Pritchard, "Collector-base impedance of a junction transistor," Proc. IRE, vol. 41, p. 1060; August, 1953; also R. L. Pritchard (reference 1); pp. 798–799. Explicitly,

$$C_d \approx (I_e w/D_b)(\partial w/\partial E_c),$$

where w is the base thickness, $(\partial w/\partial E_c)$ is the rate of change of base thickness with collector voltage, and D_b is the diffusion constant for minority carriers in the base. Alternatively, that part of the output conductance g_{22} which is due to stored charge may be written as $0.8\,\omega_c C_d = g_{c\beta}/(1-\beta_0)$, where $g_{c\beta}$ is that part of the Early collector-base conductance g_c due to the transport function β, and β_0 is the low-frequency value of β. If emitter efficiency γ is completely negligible, as assumed here, $g_{c\beta}/(1-\beta_0) = g_c/(1-\alpha_0)$; this value can be obtained easily from low-frequency parameter measurements. However, in a practical transistor, with α_0 close to unity, $g_{c\beta}/(1-\beta_0)$ may *not* equal $g_c/(1-\alpha_0)$, even though γ is negligible at *higher* frequencies.

[17] R. L. Pritchard, reference 1, p. 799. Note, however, that high-level injection *effects* were discussed first by Webster (*op. cit.*).

[18] Note that if internal tetrode biasing occurs, as it will for large values of I_e, the active cross-section area of the transistor is reduced while C_c remains fixed. Hence, high-level injection may occur for values of (C_d/C_c) considerably *less* than unity.

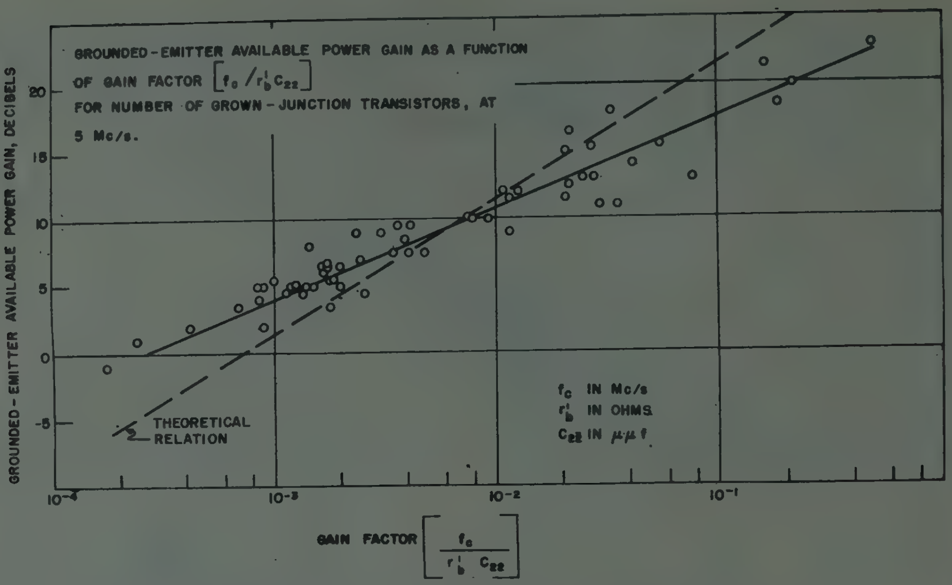

Fig. 4—Grounded-emitter maximum available power gain with resistive generator impedance as a function of gain factor $[f_c/r_b'C_{22}]$; experimental results for a number of grown-junction transistors at 5 mc.

a particular transistor in a grounded-emitter configuration, driven by a generator having a variable but purely resistive internal impedance, while the value of the abscissa is calculated from measured values of f_c, r_b' and C_{22} for that transistor. The solid line in Fig. 3 represents an "average curve" about which the experimental points appear to group themselves.

An arbitrary selection of transistors was employed for these measurements, although certain types of units specifically were excluded. The latter included transistors having poor low-frequency characteristics, transistors having alpha-cutoff frequencies <2.5 mc [in view of the 5 mc measuring frequency and of the restriction on (10)], transistors with emitter efficiencies γ appreciably different from unity at high frequencies ("slow-drool" type of alpha-frequency behavior),[19] and transistors having a large emitter-base overlap capacity.[20]

The dotted line shown in Fig. 3 represents the theoretical result calculated from (10) for $I_e=1$ ma ($r_e'=25$). Note that the solid line drawn through the experimental results lies about 3.5 db below the dotted line. Also note that no single point approaches the theoretical dotted curve more closely than about 1 db.[15]

The *same* experimental results also have been plotted as a function of the *older* parameter $(f_c/r_b'C_{22})$ derived for transistors having constant base-spreading resistance, as shown in Fig. 4. Although there definitely is a correlation between G_{av} and this parameter, as indicated by the solid line, the relationship is *not* according to theory, as shown by the dotted line calculated from (8). Also note that a large number of experimental points lie *above* dotted curve in Fig. 4, i.e., for a large number of transistors (mostly those having lower gain) measured gain exceeded theoretical gain as calculated from equation (8).

Variation of available power gain also has been measured as a function of frequency, of dc emitter current, and of dc collector voltage for a smaller number of transistors. In general, the experimental results are in agreement with the theory. For example, G varied as $\omega^{-1.5}$ to $\omega^{-1.7}$. Also, gain increased with increasing emitter current at first and attained a maximum with respect to I_e at a moderate value of I_e (of the order of 1–10 ma for the transistors measured). Finally, G increased with increasing collector voltage until collector-breakdown voltage was approached.[21]

Design Criteria for Distributed Model

Since the high-frequency power gain of a distributed model (grown-junction) transistor depends upon the same circuit parameters as does the constant-r_b' model (fused-junction) transistor, the same *general* design criteria that have been described previously[22] also may be applied to the grown-junction transistor. However, since the nature of the dependences is different for the two types of models, a few remarks together with numerical examples concerning optimum design for high-frequency performance may be of some interest.

[19] R. L. Pritchard, reference 1, p. 788.

[20] In such cases, α-cutoff frequency would be quite dependent upon dc emitter current, and the short-circuit input impedance would be largely capacitive-reactive in nature. See R. L. Pritchard, "Effect of base-contact overlap and parasitic capacities on small-signal parameters of junction transistors," Proc. IRE, vol. 43, p. 39; January, 1955.

[21] The increase of G with increasing collector voltage is due to an increase in f_c as base width is reduced by space-charge-layer widening and to a decrease in C_c.

[22] See, for example, Early, reference 2.

In addition to the different manner in which base spreading resistance influences gain in the grown-junction transistor [cf. (10) and (8)], this type of transistor also may have a graded collector-base junction, resulting in potentially lower collector-base capacitance than in the fused-junction transistor. (Generally a rather abrupt emitter-base junction is preferred in order to yield good high-frequency emitter efficiency.) The value of the collector-junction gradient represents one additional parameter for the design of high-frequency grown-junction transistors. In general, however, the gradient will be a function of majority-carrier concentration in the base, i.e., of base resistivity. For example, in one case which has been considered for computation and has been termed *proportionately graded*, the majority-carrier concentration $(N_a - N_d)$ is assumed to vary with position x through the base region as shown in Fig. 5. Consequently, the concentration gradient $A = 2N_b/w_0$, where N_b is the maximum value of $(N_a - N_d)$ in the base, and w_0 would be the base width in the *absence* of a collector depletion layer.

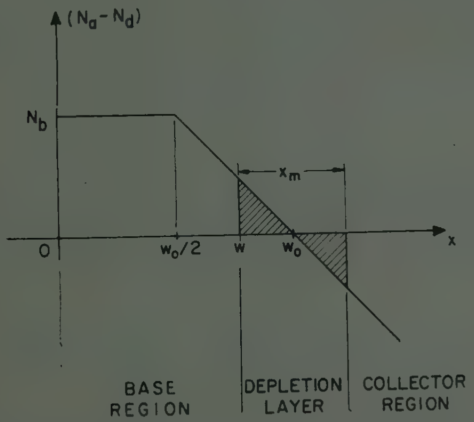

Fig. 5—Variation of majority-carrier concentration through base region of an *n-p-n* transistor having proportionately graded collector junction and abrupt emitter junction.

In order to calculate theoretical gain for a physical model of a grown-junction transistor, the circuit parameters of (10) may be related to the various physical parameters such as base dimensions, base resistivity, etc. All of these calculations are straightforward and will not be repeated here because of space limitations. If this is done, the following expression results for the first-order gain of a theoretical distributed-model transistor having substantially constant majority-carrier concentration N_b in the base:

$$G_{av} \doteq \left(\frac{0.5 q_e^{1/2}/\epsilon_0\epsilon_r}{\omega^{3/2}}\right)\left[\frac{(d/h)^{1/2} I_e^{1/2}}{S}\right] \cdot \left[\frac{(\mu_n \mu_p N_b)^{1/2} x_m}{w^{1/2}}\right]. \qquad (12)$$

A corresponding expression for the constant-r_b'-model transistor could be obtained from (8) rather than (10). Such a result already has been given by Giacoletto.[23] In (12), q_e is the electronic charge, ϵ_0 is the free-space dielectric constant (8.85×10^{-12} fd/meter), ϵ_r is the relative dielectric constant of the semiconductor (e.g., $\epsilon_r = 16$ for Ge), d is the base depth, h is the transverse height of the base (for a tetrode transistor, h would be the base-base dimension), S is the base cross-section area ($S = dh$), μ_n and μ_p are respectively the mobilities of electrons and holes in the p-type base region (which are functions of N_b),[24] and x_m is the width of the collector-base-junction depletion layer.[25]

With the help of (12), plus consideration of second-order effects, the variation of theoretical gain with each of the physical parameters for the distributed-model transistor could be discussed in detail. A few of the more pertinent results are as follows:

1. For a fixed bias and fixed cross-section geometry (of conventional values), the theoretical gain is essentially independent of base width and base resistivity (within suitable limitations) and is essentially constant to within 3–6 db, depending upon the nature of the collector-base junction.

2. For a fixed cross-section area S, theoretical gain is directly proportional to the square root of the ratio d/h of base depth to base height. Hence, for maximum gain, a long thin cross-section geometry should be employed with a line-type of base contact on the long side (dimension d) of the base. However, in practice, in order to obtain a good line contact, d/h probably would be limited to a moderate value, e.g., 2–5.

3. As cross-section area is reduced, first-order gain is increased directly. However, the second-order term C_d/C_c increases owing to an increased current density in the base region, while the second-order term involving $\omega_c C_T r_e'$ decreases owing to a decrease in total barrier capacity. *Reduction in cross-section area represents the most significant method of increasing gain over any appreciable range.*

[23] L. J. Giacoletto, "Comparative high-frequency operation of junction transistors made of different semiconductor materials," *RCA Rev.*, vol. 16, p. 37, Eq. (2); March, 1955. Note that because of the differences between this equation and (12) above, Giacoletto's semiconductor figure of merit $(\mu_n\mu_p/\epsilon_r^{1/2})$ is not directly applicable to the case of high-frequency grown-junction transistors. The corresponding figure of merit for a high-frequency grown-junction transistor having a line-type of base contact, with a linearly graded collector junction is $[(\mu_n\mu_p)^{1/2}/\epsilon_r^{2/3}]$. Note that in this case the comparison between silicon and germanium is much less unfavorable than in the case of the fused-junction transistor. (Ratio of figure of merit of germanium to that of silicon is 2.9 for case of grown-junction and is 10.7 for case of fused-junction.)

[24] See, for example, M. B. Prince, "Drift mobilities in semiconductors, I. Germanium," *Phys. Rev.*, vol. 92, pp. 681–687; November 1, 1953; Part II. Silicon, *Phys. Rev.*, vol. 93, pp. 1204–1206; March 15, 1954.

[25] For an essentially abrupt collector-base junction with collector resistivity very much less than the base resistivity, $x_m = (2\epsilon_0\epsilon_r E_c/q_e N_b)^{1/2}$, just as in the case of the fused-junction transistor. For the proportionally graded junction shown in Fig. 5, $x_m = (6\epsilon_0\epsilon_r E_c w_0/q_e N_b)^{1/3}$.

A few numerical values of gain which have been calculated for various physical parameters[26] are shown below in Table I. Values of gain are quoted at 5 mc, but

TABLE I

THEORETICAL POWER GAIN, DISTRIBUTED-MODEL TRANSISTOR

$E_e=4.5$ v; $f=5$ mc.

$c-b$ junction	ρ_b ohm-cm	w_0 mils (10^{-3} inches)	d	h	I_e ma	First-order gain db	Second-order gain db
prop. gr.	1	0.39	10	10	1	18.9	18.2
prop. gr.	8	0.39	10	10	1	18.7	17.7
abrupt	1	0.39	10	10	1	14.5	14.0
prop. gr.	1	0.39	10	2.5	1	27.9	26.9
prop. gr.	1	0.39	10	2.5	4	30.9	28.1
abrupt	1	0.39	10	2.5	4	26.5	25.0
prop. gr.	1	0.2	10	2.5	4	31.8	30.7

gain at other frequencies may be calculated from these values by adding or subtracting 4.5 db per octave change in frequency. Note, however, that the results are valid at high frequencies only if $(\omega/\omega_c)<2$. Also, it should be emphasized that these numerical values were calculated for a *conventional* grown-junction triode, as distinguished from a *p-n-i-p* or *n-p-i-n* transistor[2] or from a "drift transistor."[27]

CONCLUSIONS

The subject of maximum gain available from a junction transistor at high frequencies has been discussed briefly. A theoretical expression has been presented for calculating maximum gain available from a transistor driven by a generator having a purely resistive internal impedance in terms of the four-pole small-signal parameters of the transistor. Under certain conditions, this expression reduces to that given by Linvill for the maximum possible gain available from a transistor.

The four-pole parameters of a theoretical model of a junction transistor then were substituted in this expression to obtain a simple concise result for evaluating high-frequency power gain in terms of three fundamental transistor parameters, viz., α-cutoff frequency, medium-frequency base spreading resistance, and collector-base capacity. Two different expressions are given, one for the fused-junction type of transistor, for which a constant base spreading resistance is valid, and one for the grown-junction type of transistor, for which base spreading resistance becomes complex and frequency-dependent at high frequencies. Experimental results given for approximately 60 grown-junction transistors tend to confirm the validity of this second result, which is new.

By further relating the three fundamental high-frequency transistor parameters to physical parameters, such as base resistivity and geometry, an expression has been presented for calculating typical available power gains for the grown-junction type of transistor. It has been emphasized that the most significant method of increasing high-frequency power gain over any appreciable range is to employ a rectangular cross section with small transverse dimension and to reduce total cross-section area. For square cross-section area of 100 mil², gains of the order of 15–18 db should be available at 5 mc. Gain at other frequencies can be calculated by addition or subtraction of 4.5 db per octave frequency change. By reducing cross-section area to, say, 25 mil² and by employing rectangular cross section with a line type of contact along the longer cross section dimension, 30 db of gain should be available at 5 mc, and it should be possible to obtain oscillations from such triode transistors up to several hundred mc.

APPENDIX A

CALCULATION OF MAXIMUM AVAILABLE POWER GAIN

Available power gain for a quadripole in a circuit as shown in Fig. 1 in general is given by the equation[28]

$$G_{av} = \frac{4R_gG_L\,|\,h_{21}\,|^2}{|(h_{11}+Z_g)(h_{22}+Y_L)-h_{12}h_{21}|^2}, \qquad (13)$$

where R_g and G_L are the real components of the generator and load immittances Z_g and Y_L respectively. (For this calculation, $Z_g\equiv R_g$.) By separating the complex four-pole parameters h_{ij} into real and imaginary parts,

$$h_{11}\equiv r_{11}+jx_{11}, \qquad h_{22}\equiv g_{22}+jb_{22}$$
$$h_{12}h_{21}\equiv H_r+jH_i \qquad Y_L\equiv G_L+jB_L,$$

and by defining dimensionless variables

$$x\equiv 1+(R_g/r_{11}),\quad y\equiv 1+(G_L/g_{22}),\quad z\equiv 1+(B_L/b_{22}),$$
$$A\equiv r_{11}g_{22},\quad B\equiv x_{11}b_{22},\quad C\equiv x_{11}g_{22},\quad D\equiv r_{11}b_{22}.$$

Eq. (13) may be written in the form

$$G_{av} = \frac{4A\,|\,h_{21}\,|^2(x-1)(y-1)}{X^2+Y^2}, \qquad (14)$$

where $X\equiv(Axy-Bz-H_r)$ and $Y\equiv(Cy+Dxz-H_i)$.

In order to maximize available power gain, (14) may be differentiated with respect to each of the three variables x, y, z, corresponding to varying R_g, G_L, and B_L respectively, and the results may be set equal to zero. This yields the following three equations:

$$\left.\begin{aligned} 2(x-1)[AXy+DYz] &= X^2+Y^2,\\ 2(y-1)[AXx+CY] &= X^2+Y^2,\\ BX &= DYx. \end{aligned}\right\} \qquad (15)$$

A solution to the last two equations of (15) is

$$X = 2(y-1)Ax, \qquad Y = 2(y-1)C, \qquad (16)$$

[26] In these calculations, dc base current is assumed to be negligible so that no internal tetrode biasing exists (see reference 14). Also for calculation of emitter-base barrier capacitance, a fixed value of total emitter-base potential (built-in potential plus applied voltage) of 0.25 v was assumed. Actually, this potential depends upon injection level and upon emitter resistivity, but the dependence is slight, e.g., see Early, reference 1, p. 1286.

[27] H. Krömer, "Zur Theorie des Diffusions- und des Drifttransistors," *Arch. elekt. Übertragung*, vol. 8, pp. 223–228, 363–369, 449–504; May, August, November, 1954.

[28] For example, R. L. Pritchard, "Small-signal parameters for transistors," *Elec. Engrg.*, vol. 73, p. 903; October, 1954.

valid for *any* x, i.e., for any R_g. Physically, the conditions expressed by (16) correspond to setting the load admittance equal to the complex conjugate of the transistor output admittance, for any arbitrary R_g, i.e.,

$$\left.\begin{aligned}(G_L/g_{22}) &= (y-1) = 1 - \frac{AH_r x + CH_i}{A^2x^2+C^2},\\ -(B_L/b_{22}) &= (1-z) = 1 - \frac{DH_i x - BH_r}{D^2x^2+B^2}.\end{aligned}\right\} \quad (17)$$

Substituting (16) and (17) in the first part of (15), then yields a fourth-degree equation in x which can be factored into two quadratic equations. One of these yields no real solution for x; the other is

$$A^2x^2 - 2A^2x + AH_r - C^2 + CH_i = 0. \quad (18)$$

Alternatively, (16) and (17) may be substituted directly back into (14) to obtain an expression for G_{av} as a function of x alone. This expression then may be maximized with respect to x. This is a far simpler procedure and leads to the same result (18). Eq. (18) may be solved for

$$(x-1)^2 = (R_g/r_{11})^2 = 1 - (H_r/A) + (C/A)^2 + (CH_i/A^2), \quad (19)$$

which is (3) in slightly different form.

Then, substitution of (16) and the first part of (17) back into (14), plus manipulation of the result with the help of (16), yields

$$G_{av} = |h_{21}|^2/(2Ax - H_r),$$

where x is given by (19). This is (1) in slightly different form.

Appendix B

h-Parameter Representation of Junction Transistor in Grounded-Emitter Configuration

The grounded-emitter $h^{(e)}$ parameters for an ideal one-dimensional transistor may be calculated by conventional transformation from the grounded-emitter *admittance* or $y^{(e)}$ parameters. The latter parameters can be calculated in terms of the grounded-base y parameters[1] from the relation

$$[y^{(e)}] = \begin{bmatrix} y_\Sigma & -(y_{12}+y_{22}) \\ -(y_{21}+y_{22}) & y_{22}\end{bmatrix}, \quad (20)$$

where

$$y_\Sigma \equiv (y_{11}+y_{12}+y_{21}+y_{22}).$$

Then,[29]

$$[h^{(e)}] = \begin{bmatrix} 1/y_\Sigma & (y_{12}+y_{22})/y_\Sigma \\ -(y_{21}+y_{22})/y_\Sigma & (y_{11}y_{22}-y_{12}y_{21})/y_\Sigma\end{bmatrix}. \quad (21)$$

It should be noted that emitter-base and collector-base barrier capacities C_e, C_c are included here in y_{11} and y_{22}, respectively.

Each $h^{(e)}$ parameter can be calculated in terms of base thickness, diffusion constants, etc., if desired. However, it may be of more practical interest here to express each $h^{(e)}$ parameter in terms of appropriate low-frequency parameters and a single frequency variable, namely, ω/ω_c, or (radian) frequency ω relative to α-cutoff frequency ω_c. Thus, for the special case of unity emitter efficiency γ,[10] the current-amplification factor $\alpha=\beta$, where

$$\beta = \beta_0 \operatorname{sech}(j\omega\tau_D)^{1/2}, \quad (22)$$

is the transport factor. In (22), β_0 is the low-frequency value of β, and

$$\tau_D = 2.43/\omega_c, \quad (23)$$

where $\omega_c/2\pi$ is the α-cutoff frequency. In this special case,[30]

$$y_\Sigma = y_e'(1-\beta)(1+\mu_{ec}) + j\omega(C_e + C_c), \quad (24)$$

where

$$y_e' \equiv (1/r_e')[(j\omega\tau_D)^{1/2}\coth(j\omega\tau_D)^{1/2}] \quad (25)$$

is the emitter-base diffusion admittance, and $r_e' = (kT/q_eI_e)$ is the Shockley, *et al*, emitter resistance; μ_{ec} is the voltage-feedback factor described by Early,[11] and C_e, C_c are, respectively, emitter-base and collector-base *barrier* capacities. Similarly,[31]

$$\left.\begin{aligned}(y_{12}+y_{22}) &= \mu_{ec}y_e'(1-\beta) + j\omega C_c\\ -(y_{21}+y_{22}) &= y_e'(\beta-\mu_{ec}) - j\omega C_c\\ (y_{11}y_{22}-y_{12}y_{21}) &= y_e'\Big\{j\omega(C_c+\mu_{ec}C_e)\\ &\quad + \frac{g_{c\beta}}{2(1-\beta_0)}[(y_e'r_e')(1-\beta^2)]\Big\}\end{aligned}\right\}, \quad (26)$$

where $g_{c\beta}$ is the low-frequency collector-base Early conductance (for $\gamma=1$).

In general, μ_{ec} is very small and may be neglected relative to unity and to β. In this case, expressions for the $h^{(e)}$ parameters of the ideal transistor may be obtained as follows:

$$h_{11}^{(e)} = \{y_e'(1-\beta)[1+\epsilon]\}^{-1}, \quad (27)$$

$$h_{12}^{(e)} = [\mu_{ec} + j\omega C_c/y_e'(1-\beta)]/[1+\epsilon], \quad (28)$$

$$h_{21}^{(e)} = [\beta/(1-\beta)][1 - j\omega C_c/y_e'\beta]/[1+\epsilon], \quad (29)$$

$$h_{22}^{(e)} = \left[\frac{j\omega C_c}{(1-\beta)} + \frac{g_{c\beta}}{2(1-\beta_0)}(y_e'r_e')(1+\beta)\right]\Big/[1+\epsilon] \quad (30)$$

[29] See for example, J. S. Brown and F. D. Bennett, "The application of matrices to vacuum-tube circuits," Proc. IRE, vol. 36, p. 852; July, 1948.
R. L. Pritchard, reference 28, p. 905.

[30] If emitter efficiency is not unity, a diffusion admittance term of the form $g(1+j\omega\tau_e)^{1/2}$ described by Shockley must be added to the right-hand side of (24) and to y_{11} in calculating $h^{(e)}_{22}$. See W. Shockley, "The theory of p-n junctions in semiconductors and p-n junction transistors," *Bell Sys. Tech. Jour.*, vol. 28, pp. 449–450; July, 1949.

[31] In deriving the third of these results, use was made of the identity $\tanh Z = \coth Z(1-\operatorname{sech}^2 Z)$.

where

$$[1+\epsilon] \equiv [1 + j\omega(C_e + C_c)/y_e'(1-\beta)]. \quad (31)$$

Alternatively, for $h_{22}^{(e)}$ at high frequencies it is useful to express the second term involving the space-charge-layer widening factor in terms of the collector-base diffusion capacitance C_d due to charge stored in the base by dc emitter current.[16] The relation for the *theoretical* model is simply

$$g_{c\beta}/(1-\beta_0) = (\omega_c C_d/1.2). \quad (32)$$

However, it should be pointed out that for a *practical* transistor, the factors $g_{c\beta}$ and β_0 in general will *not* correspond to measured collector-base conductance g_c and current-amplification factor α_0 respectively.[11,16] The difference arises from the fact that although emitter efficiency γ may be *essentially* equal to unity without modifying the high-frequency behavior of the transistor, if β_0 is close to unity, $(1-\gamma_0)\neq 0$ may be comparable with $(1-\beta_0)$. In this case, in (30) for $h_{22}^{(e)}$, $g_{c\beta}/(1-\beta_0)$ should be replaced by $g_c/(1-\alpha_0)$ *for low frequencies*; however, at high frequencies, (30) as given is correct. Since in general it is not easy to differentiate between $g_{c\beta}$ and g_c, use of (32) is recommended for high frequencies. (It might be pointed out that if $C_d \ll C_c$, then at high frequencies the second term in (30) for $h_{22}^{(e)}$ will be negligible compared to the term involving C_c.)

It is especially important to note that *all* of the $h^{(e)}$ parameters *for the ideal transistor* can be calculated completely in terms of four low-frequency parameters r_e', μ_{ec}, $g_{c\beta}$, β_0, plus two barrier capacitances C_c, C_e, and two normalized functions of frequency, $y_e'r_e'$ and (β/β_0). Each of the latter two functions depends only upon the β-cutoff frequency $\omega_c/2\pi$. Curves of these two normalized functions of frequency have been presented earlier for a fairly wide range of frequencies.[1,32] At low frequencies, each of the hyperbolic functions may be represented approximately by series expansions. Thus,

$$(\beta/\beta_0) \approx [(1 - x^2/4) - j1.2x]/(1+x^2), \quad (33)$$

$$y_e'r_e' \approx [(1+x^2/4) + j0.8x]/[1+0.06x^2], \quad (34)$$

where

$$x \equiv \omega/\omega_c$$

is the ratio of frequency to α-cutoff frequency. These expansions are accurate to order x^2 and are valid up to approximately $x=2$, to an accuracy of approximately 10–20 per cent. (Somewhat less accurate series expansions have been given by Early.[1])

With the help of these series expansions, approximate expressions may be obtained for each $h_{ij}^{(e)}$ parameter in a straightforward manner. For example, if low-frequency effects are neglected, so that $(1-\beta_0)\approx 0$, particularly simple results are obtained:

$$h_{11}^{(e)} \approx [(r_e'/6)+(\omega_c r_e'/j1.2\omega)]/[1+\epsilon], \quad (35)$$

$$h_{12}^{(e)} \approx [\mu_{ec}+0.8\omega_c C_c r_e'(1+jx/6)]/[1+\epsilon], \quad (36)$$

$$h_{21}^{(e)} \approx [-1/6+(1/j1.2x) - 0.8\omega_c C_c r_e'(1+jx/6)]/[1+\epsilon], \quad (37)$$

$$h_{22}^{(e)} \approx [0.8\omega_c C_c(1+jx)+0.8\omega_c C_d(1+j0.2x)]/[1+\epsilon], \quad (38)$$

$$[1+\epsilon] \approx [1+0.8\omega_c(C_e+C_c)r_e'(1+jx/6)]. \quad (39)$$

From these equations, together with the matrix (5), the parameters required for calculating G_{av} according to (1) and (3) can be evaluated readily. For example, for $g_{22}=Re(h_{22}^{(e)})$, from (38), to the first approximation, [see assumptions 1–4 following (5)] $g_{22}\equiv 0.8\ \omega_c C_c$, whereas to the second approximation,

$$g_{22} \approx [0.8\omega_c(C_c + C_d)]/[1+0.8\omega_c(C_e+C_c)r_e'].$$

Alternatively, the theoretical transistor can be represented by an equivalent circuit if desired. The basic circuit for h-parameter representation is shown in Fig. 6(a).[33] This two-generator circuit also can be converted to a one-generator circuit by incorporating either shunt or series feedback. An approximate circuit employing shunt feedback, which is valid if $|h_{12}|\ll 1$ and if $|h_{12}|\ll|h_{21}|$, is shown in Fig. 6(b).

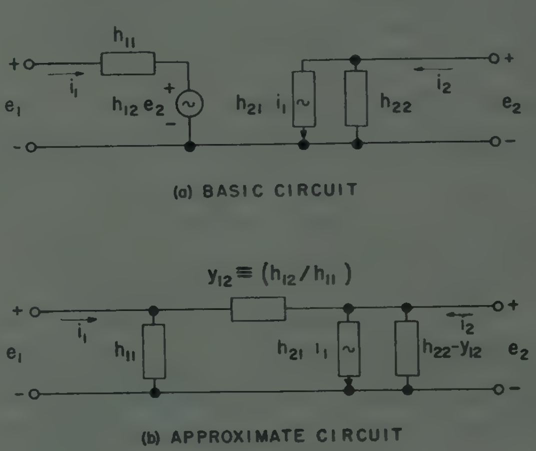

Fig. 6—Equivalent circuits for representation of a network by small-signal series-parallel h parameters; (a) Basic two-generator circuit, (b) Approximate one-generator circuit.

For grounded-emitter operation of the transistor, the elements of the approximate equivalent circuit shown in Fig. 6(b) assure reasonable and well-behaved values. For example, for frequencies less than approximately twice ω_c, relatively simple two-terminal networks for each of the elements can be constructed from the expressions given above for $h_{ij}^{(e)}$ in (35) to (38). These networks can be simplified further and reduced to simple components by neglecting certain terms with a loss of accuracy of approximately 20 per cent. Further simplification also results if it is assumed that μ_{ec} is negligible at high frequencies. The resulting circuit for *high*

[32] R. L. Pritchard, "Frequency variations of current-amplification factor for junction transistors," PROC. IRE, vol. 40, p. 1480; November, 1952.

[33] The equivalent circuits described here also have been given by the writer in "Frequency response of theoretical models of junction transistors," TRANS. IRE, vol. CT-2, No. 2, pp. 183–191; June, 1955.

frequencies is shown in Fig. 7; the base impedance z_b' also has been included in accordance with the model shown in Fig. 2(b).[34] In this circuit, the frequency variation of the current-generator constant $h_{21}^{(\epsilon)}$ has been taken into account by employing a simple *pi* network in the output circuit and by replacing $h_{21}^{(\epsilon)}$ by a constant $\alpha_0/(1-\alpha_0)$ (low-frequency value of $h_{21}^{(\epsilon)}$), independent of frequency. As a consequence of this modification, the limiting low-frequency representation of $h_{22}^{(\epsilon)}$ also can be obtained from this circuit. To obtain the low-frequency representation for h_{11} and h_{12}, additional resistances $r_\epsilon'/(1-\alpha_0)$ and $r_\epsilon'/\mu_{\epsilon c}(1-\alpha_0)$ should be shunted across the input capacitance $1.2/\omega_c r_\epsilon'$ and across C_c, respectively.

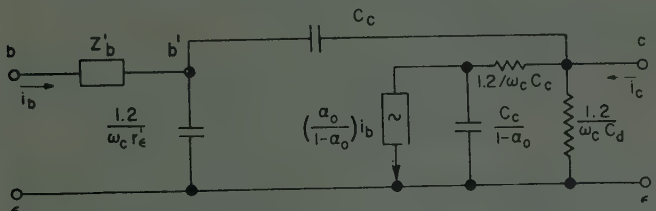

Fig. 7—Approximate high-frequency equivalent circuit for theoretical model of junction transistor; grounded-emitter operation.

On the other hand, for grounded-base operation, use of the shunt-feedback circuit shown in Fig. 6(b) leads to negative values for the admittance $(h_{22}-y_{12})$; this is not especially desirable. A series feedback circuit could be employed (e.g., see that given by Early[11] for low frequencies). However, for high frequencies in grounded-base operation, feedback is predominantly due to the effect of the base impedance, and h_{12} of the inherent transistor is negligible. Consequently, the basic circuit [Fig. 6(a)] may be employed directly for the *ideal* transistor, and at high frequencies, $h_{12}e_2$ simply may be neglected.

By combining the circuit of Fig. 6(a), sans voltage generator, with the model shown in Fig. 2(a), and by substituting appropriate low-frequency expansions for the grounded-base h_{ij} parameters of the ideal transistor, the circuit shown in Fig. 8 has been synthesized.[35] In this circuit, as in Fig. 7, a simple *pi* network has been used in the output circuit. However, in this case, although the resulting current-generator "constant" can be made independent of frequency with respect to amplitude, it is necessary to include a frequency-dependent phase shift, i.e., a constant time delay, between the input current and current applied in the *pi* network. To complete the circuit for low frequencies, the voltage generator $\mu_{\epsilon c}e_2$ and a collector-base conductance g_c must be added.[11]

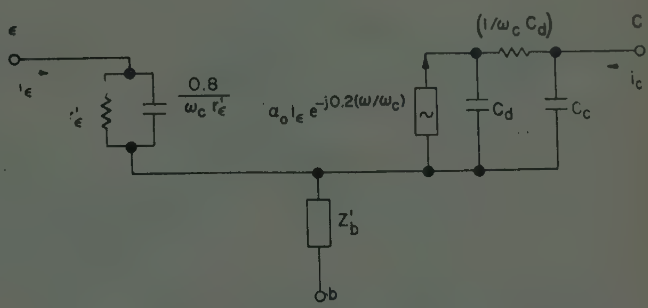

Fig. 8—Approximate high-frequency equivalent circuit for theoretical model of junction transistor; grounded-base operation.

Acknowledgment

The writer is grateful to J. Lawrence for his help in obtaining experimental data reported here and to R. N. Hall and W. N. Coffey for helpful discussions.

[34] Note the similarity between this circuit of Fig. 7 and the hybrid-*pi* equivalent circuit of Giacoletto-Johnson. (See for example, Giacoletto, reference 2, or C. W. Mueller, and J. I. Pankove, "A *p-n-p* triode alloy junction transistor for radio frequency amplification," *RCA Rev.*, vol. 14, p. 594; December, 1953, also Proc. IRE, vol. 42, p. 389; February, 1954.) The principal difference is that in the equivalent circuit above, the current generator in the output circuit is proportional to the actual input current to the over-all transistor, whereas in the Giacoletto-Johnson circuit, the current generator is proportional to a voltage at a point (b′) inside the equivalent circuit which is *not* accessible in practice. Also, in the latter circuit, the current generator is shown to be independent of frequency, whereas in theory, the output current should lag the $b'-\epsilon$ voltage by an appreciable phase angle, e.g., 22 degrees at the α-cutoff frequency.

[35] Note the simplicity of this approximate circuit relative to the more exact circuits employing RC transmission lines (Pritchard, reference 16, Fig. 1; reference 1, Fig. 13, 16. Also, J. Zawels, "Physical theory of new circuit representation for junction transistors," *Jour. Appl. Phys.*, vol. 25, p. 978; August, 1954). A simplified equivalent circuit incorporating a time delay (delay line) between input and impressed-output currents also was presented by W. F. Chow and J. J. Suran, "Transient analysis of junction transistor amplifiers," Proc. IRE, vol. 41, pp. 1126–1127; September, 1953.

IRE Standards on Radio Receivers: Method of Testing Receivers Employing Ferrite Core Loop Antennas, 1955*

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1.00. Introduction. The technique for introduction of a test signal into a loop antenna (see "Standards on Radio Receivers—Methods of Testing Amplitude-Modulation Broadcast Receivers—1948," Section 4.01.03, for example) has long been employed with receivers using air-core loop antennas. These antennas are normally of a more or less flat or "pancake" construction and, in general, lend themselves to the method described in the standard without ambiguity. When this method is extended to loop antennas wound on cores of high permeability in which the length-to-diameter ratio is high, it tends to break down. When the test loop and this type of antenna are coaxial, it is usually not feasible to assign a spacing between these two for calibration purposes. Moreover, the use of the induction field to simulate the actual radiation field received by the loop is a satisfactory procedure only if the loop is immersed in a reasonably uniform field. This is substantially the situation with flat air core loops using the aforementioned technique but is not approximated satisfactorily when the relatively long ferrite core loop antenna is employed. This present standard describes a modification of the existing techniques which allows for the measurement of a receiver employing a ferrite-core loop antenna with the same precision as that obtained in the measurement of air-core loop antennas.

* Reprints of this Standard, 55 IRE 17.S1, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy. A 20 per cent discount will be allowed for 100 or more copies mailed to one address.

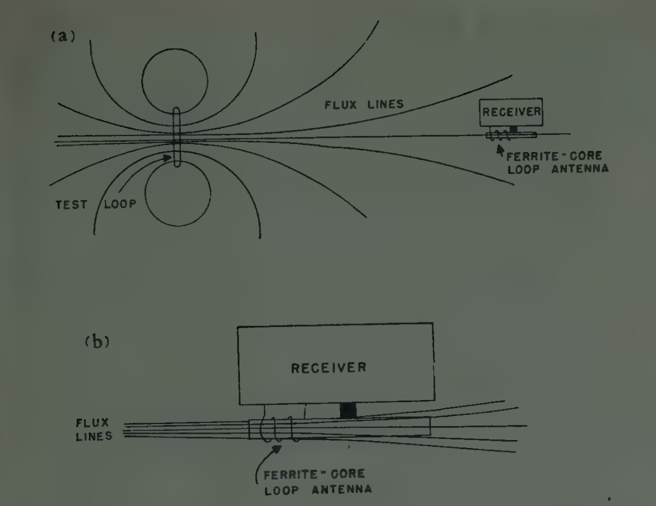

Fig. 1—Field configuration when air loop practice is followed with ferrite core loop antenna.

2.00. Apparatus Required.

2.01. Radiating Test Loop. The test loop employed can be identical to the one described in "Standards on Radio Receivers—Methods of Testing Amplitude-Modulation Broadcast Receivers—1948," Section 5.00. It should be observed, however, that the calibration relations of that section and Section 4.01.03.01 of the same standard refer to the field directed along the axis of the loop. Although this is not the field that is to be employed in this standard, the same loop design is implied.

2.02. Shielded Room (*Screen Room*). The relations presented for the field due to the loop assume free space conditions. Measurements generally are made in a screen room, and the precaution of Section 5.02 of the "Standards on Radio Receivers—Methods of Testing Amplitude-Modulation Broadcast Receivers—1948" applies. If there is any doubt concerning the validity of the measurements due to the proximity of fixed large metal objects, the field within the screen room can be compared with the field at the same distance in free space and a suitable correction factor applied to the screen room field. This correction is explained more fully in Section 3.03.

2.03. Other Equipment. The standard signal generators and receiver output measuring devices required for any particular test should have the characteristics specified in Section 3.00 of "Standards on Radio Receivers—Methods of Testing Amplitude-Modulation Broadcast Receivers—1948."

3.00. Method of Measurement. This new standard describes a method of introducing a known signal into a ferrite-core loop antenna and therefore replaces Section 4.01.03 of "Standards on Radio Receivers—Methods of Testing Amplitude-Modulation Broadcast Receivers—1948," when a receiver employing such an antenna is being measured. The remainder of the test procedures specified in Section 4.00 of that standard still apply.

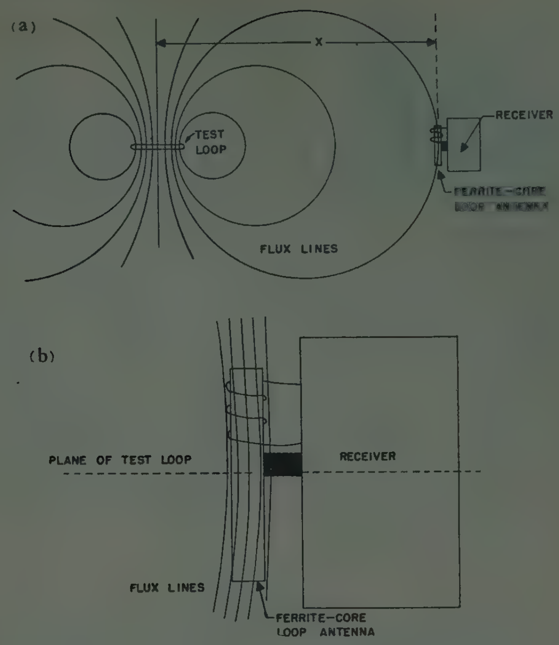

Fig. 2—Field configuration suitable for ferrite core loop antennas.

3.01. Orientation of Radiating Loop and Receiver under Test. Fig. 1 shows the magnetic field that would exist at a ferrite-core loop antenna if the orientation of radiating loop and receiver loop normally used with receivers employing air loop antennas were used (this figure and subsequent ones do not demonstrate the effect on the magnetic field configuration of the presence of the high-permeability ferrite rod). Notice that the field is not constant over the length of the rod; and since the length of the ferrite rod is generally significant compared to the spacing between radiating loop and rod, it is difficult, in general, to assign a value to the effective field at the antenna.

The recommended orientation is as shown in Fig. 2. The axis of the ferrite rod is placed normal to the plane of the test loop with the center of the rod in the plane of the test loop. The antenna rod is now located in a substantially constant field to which an effective value can be assigned.

3.02. Calibration of Effective Field. The significant component of field near the radiating test loop is a magnetic field that varies inversely as the cube of the distance from the loop. The receiving loop is essentially responsive to a magnetic field, and the magnitude of the field can be calculated in terms of the test loop parameters, the current in the test loop, and the spacing between the radiating and receiving loops. This can also be expressed in terms of the equivalent electric radiation field in volts per meter that would be accompanied by a magnetic field of the calculated value. The relation between the observed current in the coil and the equivalent electric field intensity is, to a close approximation, given by the formula:

$$E = \frac{30\pi n_1 a_1^2}{x^3} I_1,$$

where E = equivalent electric field intensity in volts per meter at the receiving loop antenna, n_1 = number of turns of radiating loop L, a_1 = radius of radiating loop L in meters, x = distance in meters between the center of radiating loop L and the axis of the ferrite rod receiving loop, I_1 = current in radiating loop L in amperes. If the loop is constructed as described in Section 5.01 of the "Standards on Radio Receivers—Methods of Testing Amplitude-Modulation Broadcast Receivers—1948," and a spacing of 24 inches (x) between the axes of the radiating test and receiving loops is employed, the above equation can be reduced to a simple relationship between the indicated output voltage of the signal generator (V) and the effective electric field strength in volts per meter as follows:

$$E = 0.05V.$$

In other words, the field in db below one volt per meter is 26 db below the signal generator output in db below one volt.[1]

[1] This relation is approximate in that it assumes the loop is a point source. For greater accuracy nonuniformity of field over length of ferrite rod must be taken into account. The 26-db value appears to be sufficiently accurate for normal measurement purposes.

3.03. Effect of Screen Room. The equation relating effective electric field intensity to signal generator output is derived assuming that the induction field of the radiating loop varies inversely as the cube of the distance from the loop as occurs in free space. If the screen room is not sufficiently large, reflections from its walls will tend to modify the field from its ideal value. It is useful to set up a receiver, radiating loop, and signal generator in "free space" (as far from large metal objects as possible) in the manner prescribed in this standard, and adjusted to a frequency that can be used outside the screen room without undue interference. The signal generator output required to produce a suitable reference output from the receiver is recorded. The entire set-up is then transferred to the screen room, and the signal generator output readjusted, if necessary, to produce the same reference output from the receiver. This change in signal generator output is a measure of the distortion of the field configuration due to the screen room. The correction should be used to modify the relation between effective field strength and signal generator output given in Section 3.02. Since the calibration in general is not independent of frequency, it should be made at or near the frequency of interest. If the locations of the instruments within the screen room are modified, a new calibration may be required.

A Microwave Phase Contour Plotter*

J. S. AJIOKA†

Summary—**A simple, easily built rf phase contour plotting device is described. This phase plotter differs from the conventional plotter in that two field-sampling probes are used instead of one. Instead of a fixed reference signal taken directly from the source, a phase reference is taken from the field itself. This technique offers several practical advantages over conventional methods.**

THE MEASUREMENT of phase in the field of microwave radiators is important in the design of antenna components and in the study of the effects of objects placed in the field. The determination of equiphase contours and the location of phase centers of feeds for reflectors and lenses are of much value in the design of such antennas. Phase measurements are also important in the study of the effects of weatherizing covers and radomes.

This paper describes a simple, inexpensive, easily made device for manual phase contour plotting, with primary application to locating the phase centers of primary radiators. It has some advantages over the conventional phase contour plotting apparatus and is comparable in accuracy. This device is not intended to replace some of the rather complex and expensive automatic phase plotters[1] now in existence, but rather to afford a design for a simple inexpensive phase plotter that can be built in a relatively short time.

In the conventional method for measuring phase, a sample of energy from the field is picked up by a probe and compared in phase with a reference signal which comes directly from the source.[2] The phase of one signal is varied with respect to the other to produce an interference resulting in a maximum or minimum. The amount of phase shift is a measure of the difference in phase between the field sample and the reference. A slotted line with the reference signal inserted through the traveling probe is often used as the phase shifter and mixer.

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† Hughes Res. Labs., Hughes Aircraft Co., Culver City, Calif.

[1] R. M. Barrett and M. H. Barnes, "Automatic antenna wavefront plotter," *Electronics*, vol. 25, pp. 120–125; January, 1952.
[2] S. Silver, "Microwave Antenna Theory and Design," McGraw-Hill Book Co., Inc., New York, N.Y., ch. 15; 1949.

The conventional phase measuring apparatus has several practical disadvantages which are as follows:

1. The necessary rf rotating joints or flexible cables complicate the rf plumbing; and unless much care is taken, may cause phase shift with movement.
2. If there is a wide variation in the amplitude of the field whose phase is to be measured, the minima readings in the mixer cannot all be made sharp by a simple interference procedure.[3]
3. All components in the rf system must be well matched or false minima can occur.[2]
4. Phase contour plot, usually not direct, must be plotted from data taken from phase shifter readings.
5. The accuracy is impaired by small changes in frequency during the time for making a complete plot.[2]

A method using two field-sampling probes instead of one eliminates the above disadvantages. Instead of a fixed reference directly from source, a phase reference is taken from the field itself. Fig. 1 is a photograph of this phase plotter. The stand in the background is an instrument for projecting the horn onto plotting paper.

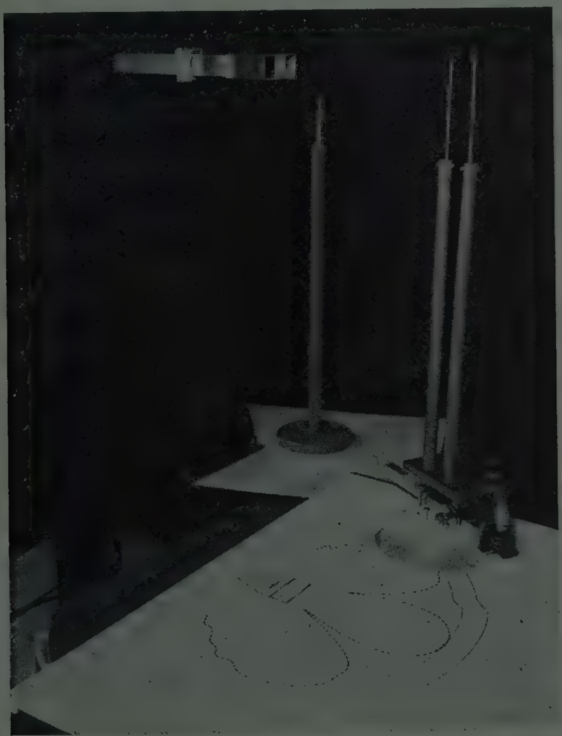

Fig. 1—Phase contour plotter in operation.

The device (See Fig. 2) is described as follows:

Two probes spaced a fixed distance apart are connected respectively to the E and H arms of a magic tee where their signals are mixed. A detector is placed on one of the remaining arms and a matched load is placed on the other. The connection is such that in-phase signals from the probes will destructively interfere at the detector to give a minimum. Since the probes are close enough together so that they are usually in a field of approximately the same strength, the minimum is sharp. Since the magic tee with the detector is an integral part of the two-probe combination, there is only an audio cable that need be flexible.

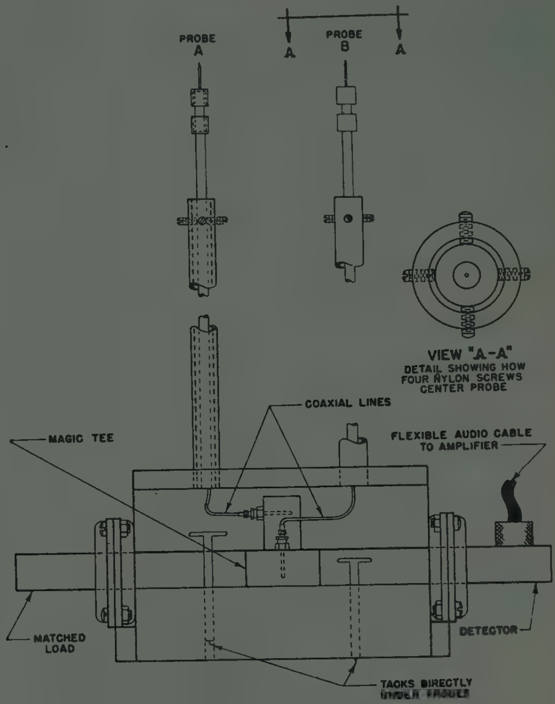

Fig. 2—Sketch of phase contour plotter.

To plot equiphase contours, the probe assembly is placed in the field of the radiator whose phase fronts are to be plotted. The probe assembly is placed over a flat surface on which the phase contour is plotted. The flat surface is $\frac{1}{2}$ inch plate glass covered with heavy paper.

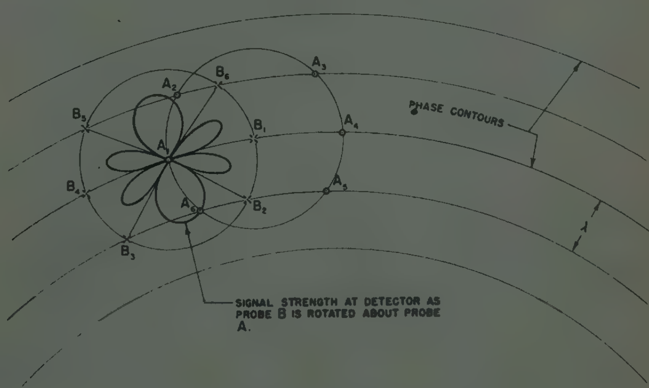

Fig. 3—Diagram illustrating the procedure of plotting phase contours.

(See Fig. 3). The tack under probe A is pushed into the surface (paper) and the probe assembly is rotated about probe A and the positions of minimum readings are

[3] With somewhat more complication, this difficulty can be eliminated. See F. L. Vernon, Jr., "Application of the microwave homodyne," Trans. IRE, vol. AP-4, pp. 110–116; December, 1952 and J. Bacon, "An automatic X-band phase plotter," Proc. NEC, Chicago, Ill.; October, 1954.

marked by pushing tack "B" into the paper. These B-tack points (marked by x's) are points of equal phase. The tack is lifted from point A and the probe assembly is rotated about one of the B-tack points. Since the center of phase is somewhere in the region of the aperture of the radiating element under test, there will, in general, be no ambiguity as to which of the B-points are on the same phase contour as point A. Of course, there would be no ambiguity if the probe spacing were less than a wavelength but the sensitivity would be decreased. These B-points are labeled B_1 and B_4 in the figure. As the assembly is rotated about point A_1, the null-points B_1 and B_4 only are considered and the center of rotation is then moved to either B_1 or B_4 (say B_1) and point A_4 on the same phase front as A_1 is found. This process of alternately rotating about probe A (circles) and probe B (x's) is repeated to plot a complete equiphase contour.

If the electrical line length from probe A to the detector is equal to that from probe B to the detector as was intended, the A and B points will be in phase. Otherwise, the points A (circles) and the points B (x's) are two different sets of equiphase points. If curves are drawn through points A and through points B, the normal separation between the curves will be the difference in electrical lengths of the lines from the two probes. This difference in line lengths can be calibrated by picking a fixed point in the field, rotating about probe A and establishing a null-point for probe B. Then place probe B at fixed point, rotate about probe B and establish a null-point for probe A. Distance between these null-points will be twice the difference in line lengths.

Since the path lengths from the probes to the detector are made to be as nearly identical as possible, the phase front plotter is quite insensitive to variations of frequency.

If the phase fronts are smooth, there is no ambiguity as to which points are on the same contour; but if there is a sudden phase change as would occur over a minimum in the amplitude pattern of the radiator under test, care must be taken so that a true phase contour would be followed. For this reason, amplitude patterns are also taken with a single probe with a center of rotation about the approximate center of phase of the radiating element. A deep minimum in the amplitude pattern will give warning that a sudden change in phase is to be expected. A typical amplitude and phase plot of such a situation is given in Fig. 4. Fig. 4 is an amplitude (dashed lines) and phase (solid lines) plot of an E-plane horn with a septum. This horn was chosen as an example because of the deep minima in its amplitude pattern. If the minimum were infinitely deep, the phase change would be a sudden step of 180 degrees. In general, the minima are not infinitely deep and the phase contour is like that of Fig. 4. To plot points about the region of rapid phase change, a partial curve is drawn in the smooth phase region and points on this curve are used as reference points for establishing points closer than the probe spacing. Thus, we creep up to the region near the amplitude minimum and finally we span the amplitude minimum to establish points on the phase contour on the other side of the amplitude minimum. With a little more complexity a probe assembly could be made so that the probe spacing could be varied. Once the probe spacing is changed, it is kept unvaried during a set of measurements. This eliminates the difficulty of the above procedure. Such a variable spacing probe assembly has not been made but a fixed spacing of approximately one and one-half wavelengths was chosen to give sharp minima readings. The probes for each polarization were of conventional design.

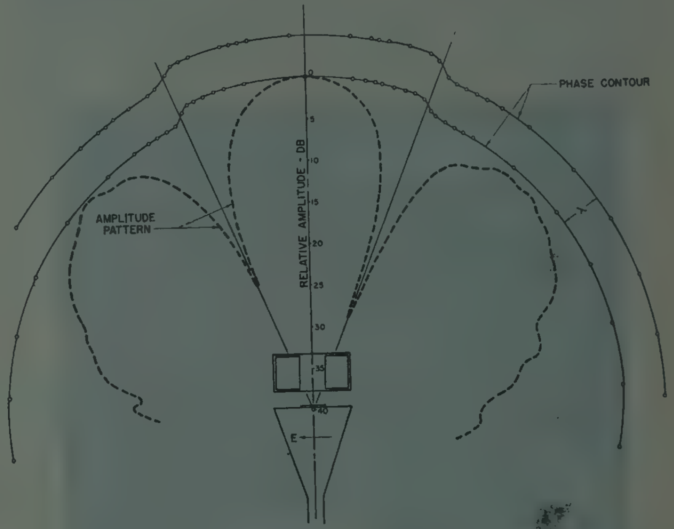

Fig. 4—Phase and amplitude plot of horn with septum.

One obvious disadvantage of the above procedure is the possibility of cumulative errors, but in the application to locating the phase centers of primary feeds this is not serious because only groups of three points on the phase contour are necessary for location of the phase center. However, the simplicity, inexpensiveness, and ease of operation make this phase plotter a practical laboratory tool.

Acknowledgment

The author gratefully acknowledges the encouragement given by all members of the Microwave Antenna Section at NEL headed by J. J. Thomas and formerly headed by E. K. Abbey. The author also wishes to thank J. H. Jensen who built the first models.

The Application of Dielectric Tuning to Panoramic Receiver Design*

T. W. BUTLER, Jr.†, ASSOCIATE, IRE, W. J. LINDSAY†, ASSOCIATE, IRE, AND L. W. ORR†, ASSOCIATE, IRE

Summary—**This paper describes a method of utilizing the voltage tuning characteristics of ferroelectric capacitors in a wide range, superheterodyne, dielectric-tuned, panoramic receiver. Continuous tuning over a 2:1 frequency band is obtained in frequency ranges up to 110 mc. Some of the problems encountered in this application are described, and a method of optimizing the parameters of specific materials is discussed. No detailed technical discussion is presented. The application of these capacitors to various types of circuitry is briefly indicated.**

Introduction

FERROELECTRIC tuning techniques utilize the nonlinear electrical characteristics of certain types of ferroelectric materials. Barium-strontium titanate materials constitute the major class of dielectrics that are presently being applied to ferroelectric tuning devices. The use of capacitors constructed from these materials as the basic tuning elements in panoramic receiver front ends is the subject of this paper.

It should be noted, however, that ferroelectric tuning techniques are being applied to sweep generators, spectrum analyzers, search receivers, and other equipments where wide-range, rapid-scan devices are useful.

This paper describes a method of utilizing the voltage tuning characteristics of ferroelectric capacitors in a wide-range superheterodyne, dielectric-tuned, panoramic receiver. This receiver employs titanate ceramic capacitors as tuning elements in the rf, mixer, and local oscillator tank circuits. The capacity of the tuning elements is varied by changing the electric field applied to the capacitor.

The receiver described is incorporated in a rather complete laboratory test unit which was designed specifically for testing dielectric-tuned receiver front end assemblies. This unit consists of an assemblage of power supplies, a display oscilloscope, and control panels. The essential components of this test unit are indicated in the block diagram of Fig. 1. The receiver front-end assemblies are plug-in units containing the electrical electrically tunable stages, i.e., the rf, mixer and local oscillator stages.

Characteristics of Capacitors

The ceramic capacitors used as the tuning elements in the front end assemblies are of the barium-strontium titanate class. These were developed in the Department of Electrical Engineering Laboratories at the University of Michigan. A commercially made body material is used, but methods of plating, connecting, and potting capacitors are still laboratory processes.

To obtain the maximum tuning range with good receiver stability and sensitivity, the capacitor body material should have the following characteristics:

1. A large change in dielectric constant, ϵ, with applied field,
2. A small temperature coefficient of ϵ over a wide temperature range, and
3. Low loss over the desired frequency range.

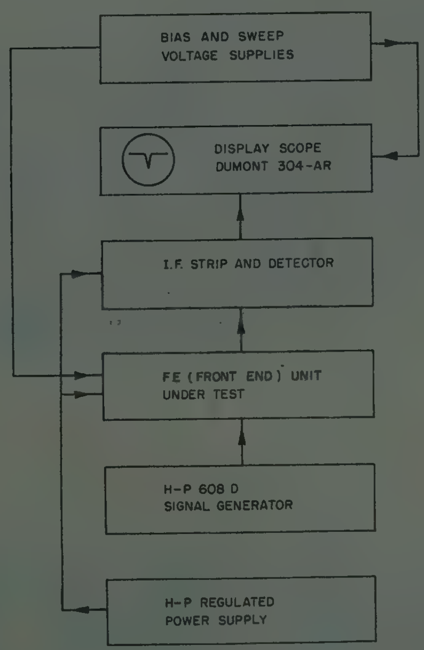

Fig. 1—Block diagram of wide-range dielectric tuned panoramic receiver.

Materials which exhibit large changes in dielectric constant with applied field also exhibit large changes in dielectric constant with changes in temperature. Conversely, materials which have a small temperature coefficient of dielectric constant over a wide temperature range generally lack field sensitivity and are not suitable for dielectric tuning. It has not been possible to obtain body material for the capacitors which possess all three of the desirable characteristics.

Fig. 2 (next page) shows an ϵ-T-E surface for a typical ferroelectric ceramic material of rather high dielectric constant. Ordinate of surface gives the dielectric con-

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† Electronic Defense group, Engrg. Res. Inst., Univ. of Mich., Ann Arbor, Mich.

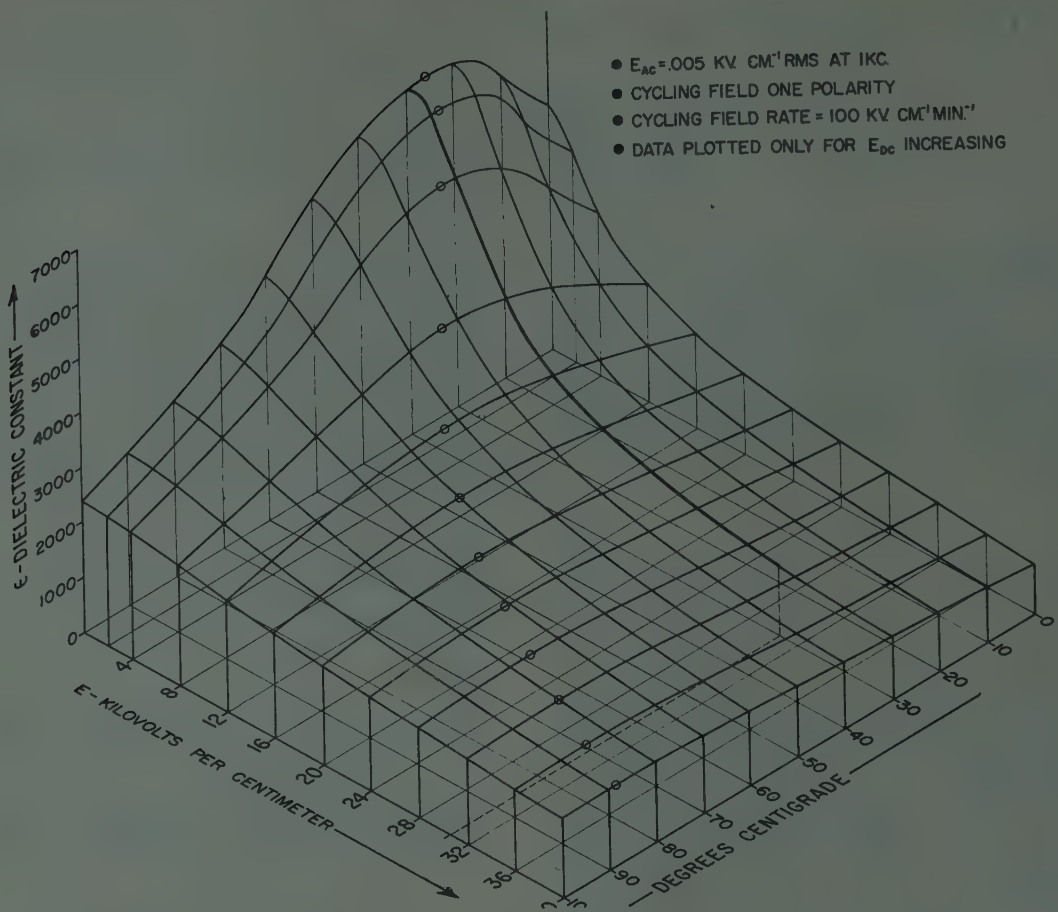


Fig. 2—ε-T-E surface for aerovox Hi-Q 41.

stant, with abscissae of temperature and dc electric field. Small circles indicate points of zero temperature coefficient. Surface representation gives a very clear display of the characteristics of the material, and its use in the design of the tuning elements is illustrated by the following numerical example.

Consider the design of the rf amplifier tank circuit which is to be tunable from 50 to 100 mc. Assuming a value of 160 $\mu\mu$f for the total tank capacitance at 50 mc, gives 40 $\mu\mu$f for the total capacitance at 100 mc. If the fixed shunt capacitance representing the tube, wiring and strays is 10 $\mu\mu$f, the tunable element must vary from 150 $\mu\mu$f to 30 $\mu\mu$f. This element will be a pair of ceramic capacitors connected in series, each having a value of 300 $\mu\mu$f at zero applied field.

The capacitance variation as the electric field is applied may be obtained from the contour curves of Fig. 2. Consider the curve for 30 degrees C. It is noted that a 5:1 variation in dielectric constant is obtained when the applied field varies from zero to 30.5 kv/cm. If the material is 0.05 cm thick, the required field is furnished by an applied voltage of 1,525 volts.

Effect of a temperature chage on tuning may be calculated by noting dielectric constant at 20 degrees C. is almost identical with that at 30 degrees C. with zero applied field. Therefore, lower frequency will still be 50 mc. At a field of 30.5 kv/cm, dielectric constant is lowered about 4 per cent below its value at 30 degrees C. The minimum capacitance value at 20 degrees C. will therefore be 28.8 $\mu\mu$f, giving an upper frequency of

$$f_2 = \left\{\frac{30 + 10}{28.8 + 10}\right\}^{1/2} \cdot 100 = 101.5 \text{ mc.}$$

This does not imply a tracking error of 1.5 mc since the oscillator circuit will also experience a similar temperature-induced change. It does, however, represent an increase in the tuning range of 3 per cent, so that as the equipment warms up from 20 degrees C. to 30 degrees C., the tuning range will shrink by this amount.

The effect of capacitor tolerance may be disposed of by noting that an increase of 10 per cent in the zero field capacitance due to a larger electrode area is accompanied by an increase of 10 per cent in the capacitance at all fields, and can therefore be compensated throughout the tuning range by an appropriate decrease (9.1 per cent) in the tuning inductance. To insure that all capacitors of the receiver have approximately the same tuning characteristic, i.e., the same ϵ-T-E surface, they may all be cut from the same wafer of ceramic material. The ceramic materials now available are sufficiently homogeneous that variations in tuning characteristic are less than 1 per cent between samples cut from the same wafer of the material.

ϵ-T-E surfaces have been obtained for a number of materials suitable for electric tuning. A survey of these surfaces is of great assistance in selecting the best available material for a particular set of conditions.

Low loss is an important factor in electric tuning, and tests show that this is not entirely a property of the ceramic material. The type and quality of metal used for electrodes, and its thickness and uniformity, all have important effects on the loss. Capacitors for the tuning units were made with 0.020-inch cubes of ceramic having one mil thick electrode plating extending to the edges of the ceramic faces. Fig. 3 shows typical capacitors made in this manner. A plastic coating over the dielectric seals out moisture, preventing excessive losses and possible electrical breakdown.

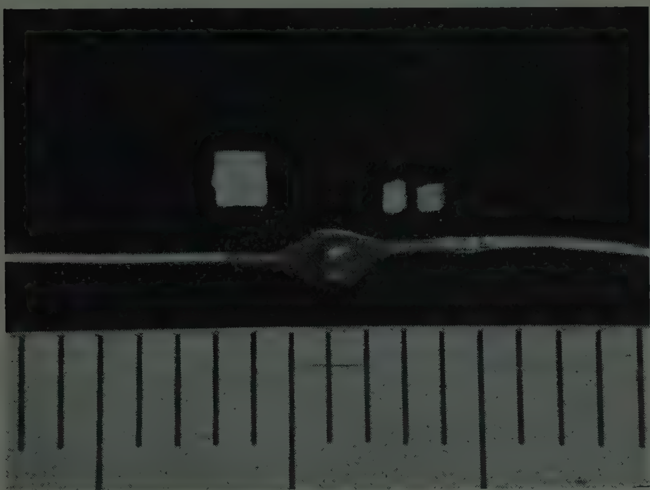


Fig. 3—Typical subminiature ferroelectric capacitor.

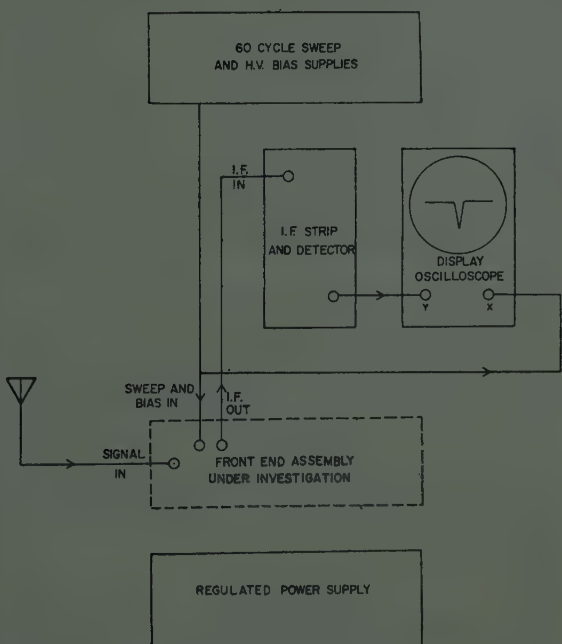


Fig. 4—Block diagram of receiver.

Receiver Design

The receiver is a single conversion superheterodyne type with a tuned rf stage, tuned mixer, and local oscillator in the tunable front end unit, and a 20 mc IF strip driving a crystal detector (Fig. 4). The three tuned circuits in the front-end unit are tuned by changing the dc bias voltage on the tuning capacitors. The receiver is swept by applying a superimposed 60 cps alternating voltage to the tuning capacitors. The dc bias voltage may be varied from 0 to 1,000 volts, and the ac sweep voltage may be varied from 0 to 1,000 volts rms. No power is absorbed from the dc bias supply in sweeping the receiver over the maximum range. The reactive power absorbed from the ac sweep voltage supply is generally less than 0.1 volt-amperes. The three tuned circuits are tracked by suitable adjustments of the dc bias and the ac sweep voltages. The IF strip is of the synchronous-tuned type[1] with a 3 db bandwidth of approximately 100 kc. The output of the crystal detector feeds the vertical input of the DuMont 304-AR oscilloscope. The horizontal input of the scope is fed a sample of the 60 cycle sweep voltage applied to the oscillator. This results in a panoramic display on the oscilloscope.

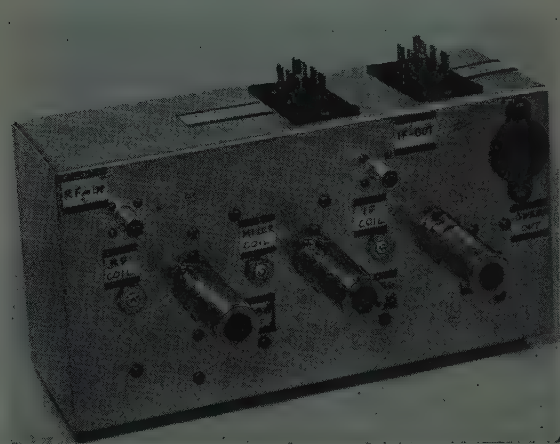


Fig. 5—Exterior view of a front-end unit.

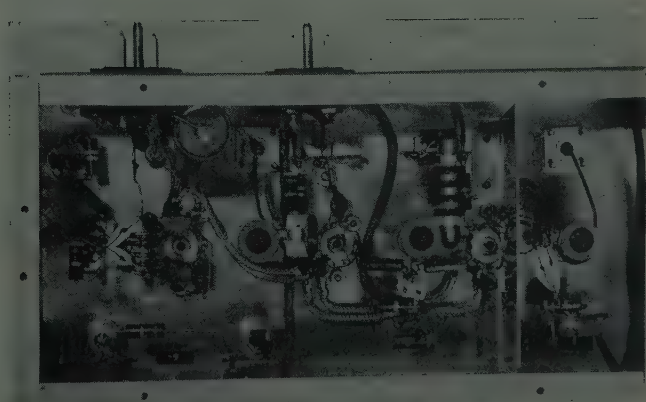


Fig. 6—Interior view of a front-end unit.

Front-End Units

Exterior and interior views of one of the early FE (Front-End) units are shown in Figs. 5 and 6. The three sets of tuning capacitors can be seen in Fig. 6. A sche-

[1] The individual stages of a synchronous-tuned IF strip are single tuned and all stages are tuned to the same frequency.

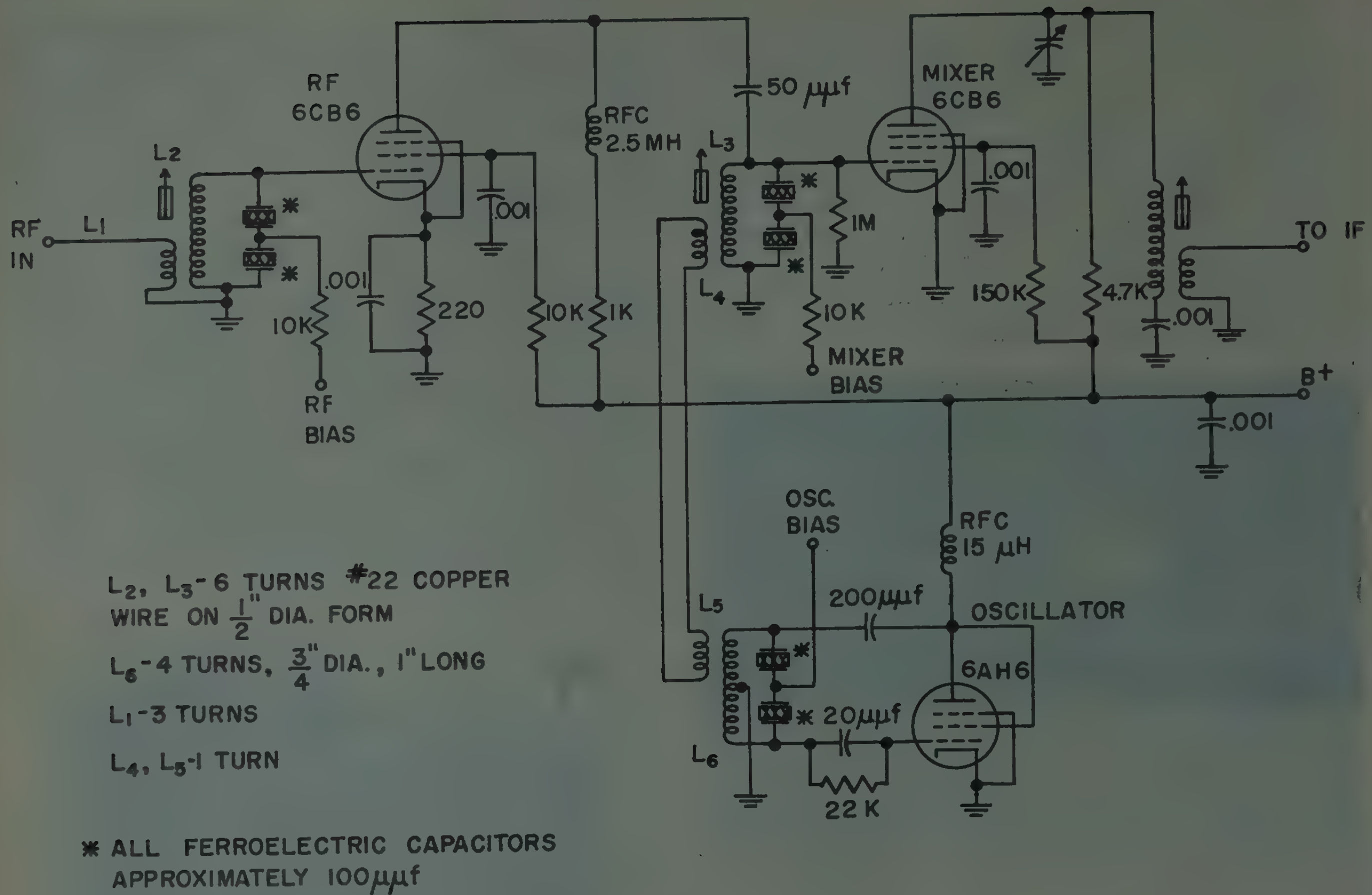


Fig. 7—Schematic of receiver front-end unit.

matic of one of the later FE units is shown in Fig. 7. The circuit design is conventional in some respects, but requires some precautions. The principal points of difference in circuitry occur in the tuned circuits and in the method of connecting the bias and sweep voltages. Capacitances to ground are kept as low as possible to reduce the effect of shunt circuit capacitance on the tuning range. This also dictates use of tubes with low shunt capacitance. The 6CB6 tube has good noise characteristics, and has performed very well in the frequency ranges considered.

A pentode mixer is used to keep the circuitry simple, and still provide good signal-to-noise ratio. Link coupling is used to obtain oscillator injection. This method of injection reduces the possibility of spurious responses due to oscillator harmonics.

The Q of the tank circuits used is lower than ordinarily encountered in receiver design. This is due to the relatively low Q of the capacitors. High g_m tubes must be used to obtain satisfactory results. The oscillator circuit is quite critical with respect to the g_m of the tube, requiring at least 7,000 to 8,000 μmhos to oscillate satisfactorily. The triode-connected 6AH6 has given good results as the oscillator tube. The 6J4 triode has also performed satisfactorily.

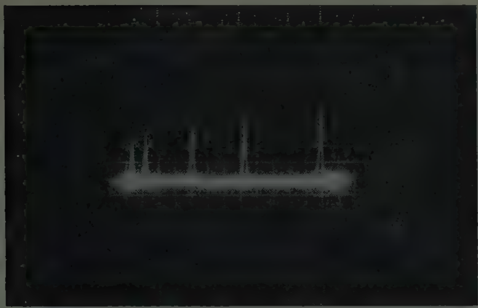
The bias and sweep voltages are decoupled from the rf circuits to avoid increases in rf loading and stray capacitance. Pairs of series-connected capacitors are used in all the tank circuits. This affords smaller minimum capacitance and at the same time supplies a convenient dc block for the bias voltage. The rf ground end of the rf and mixer coils, and the feedback point of the oscillator coil, are connected to the chassis. This permits a single connection to be used for the bias and sweep voltages, which aids in keeping stray capacitances to a minimum.

It was pointed out in an earlier section that it is desirable to use titanate materials which have a small temperature coefficient of dielectric constant over a wide temperature range. If the receiver is to be used over an extended temperature range, methods of temperature compensation will be required to maintain receiver tuning range and sensitivity. It has been found that for laboratory operation temperature compensation was not necessary.

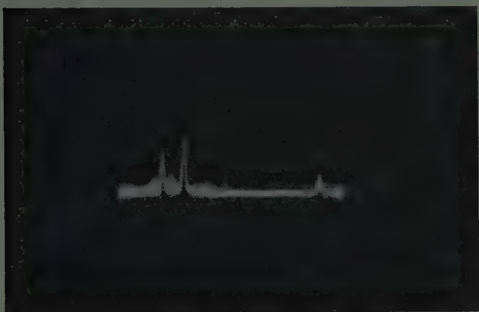
Results

To date, two working units, Models FE-2 and FE-3 have been built and tested. FE-2 tunes from 28 to 60 mc, while FE-3 tunes from 55 to 110 mc.

The sensitivity of the units may be defined as that signal level at the input of the receiver which will pro-

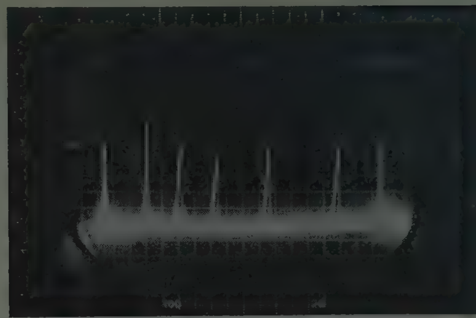


(a) Marker signals at 28, 30, 40, 50, and 60 mc.

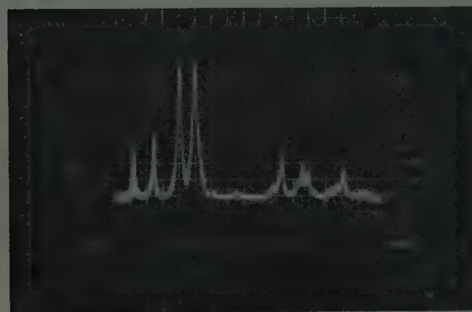


(b) Local spectrum using short wire antenna.

Fig. 8—Oscillograms showing response of FE-2.



(a) Marker signals at 55, 65, 75, 85, 95, 105, and 110 mc.



(b) Local spectrum (Detroit) with high gain discone antenna.

Fig. 9—Oscillograms showing response of FE-3.

duce a "pip" on the scope twice the apparent average height of the internal noise of the receiver.[2] The sensitivity of both units under the swept conditions listed is 5($\pm$1) microvolts into 50 ohms. When tuned to a fixed frequency, the sensitivity goes up to approximately 1.5 microvolts. The internal noise generated in the receiver is a function of the sweep width, and of other factors, the exact nature of which are not yet known. The overall bandwidth of this system is approximately that of the IF strip, or approximately 100 kc. The half-power point resolution of the receiver is approximately 150 kc. The reliability of the receiver depends primarily upon the reliability of the tuning capacitors. Recently, excellent results have been obtained in this respect. Improvements in the methods of manufacturing the capacitors from the commercially available body materials have resulted in capacitors that are expected to give continuous operation over long periods[3] with few failures.

Fig. 8 shows two oscillograms taken with FE-2 in operation. In the upper picture, marker signals of 20 microvolts at 28, 30, 40, 40 and 60 mc indicate the frequency spectrum. The lower picture shows the local strong signal spectrum with a short piece of wire used as an antenna. Fig. 9 shows a similar display for FE-3. The upper picture contains markers at 55, 65, 75, 85, 95, 105, 110 mc. The lower picture was made with a high-gain antenna oriented toward two TV stations, 30 miles distant. At the lower end of the spectrum are the audio and video signals of the two TV channels, while in the higher regions numerous FM radio stations may be observed. The frequency scale distortion noted in the upper picture is due to the nonlinear characteristic of the tuning capacitors.

Work is presently being done to extend the tuning range to higher frequencies. Preliminary tests on a higher frequency FE unit indicate that attaining the 200 mc region is feasible. This requires different types of circuits than those used in FE-2 and FE-3. At frequencies much above 100 mc the most satisfactory oscillator circuit is the ultra-audion.[4] Oscillators of this type have been built and tested up through 400 mc with tuning ratios of 1.5:1. There are two major problems in connection with extension of the frequency range.

1. It is difficult to manufacture capacitors with capacitances low enough to be useful in high-frequency circuits.

2. Rf losses in the dielectric increase with frequency, resulting in lowered Q of the tuned circuits.

Although the receiver described in this paper uses a relatively slow scan rate (60 cps), these are applications for which a faster scan rate may be desirable. Faster

[2] J. B. L. Foot, "Wideband VHF panoramic receiver," *Wireless World*, vol. 59, p. 392; September, 1953.

[3] At this writing the reliable life of capacitors under operating conditions cannot be evaluated since life-test results are incomplete.

[4] J. F. Reintjes and G. T. Coate, "Principles of Radar," 3rd ed., McGraw-Hill Book Co., Inc., N. Y., N. Y., p. 706; 1952.

scan rates are readily obtainable with dielectric tuned oscillators and have been reported up to 500 kc for small deviations.[5] To judge from the relaxation measurements on several samples of titanate ceramics, the upper practical limit of sweeping in a panoramic receiver appears at present to be about 100 kc.

Conclusion

There are many applications in the field of instrumentation in which dielectric tuned tank circuits could be used to advantage, i.e., sweep generators and spectrum analyzers. Dielectric-tuned, wide-range, vhf swept oscillators which may be suitable for use in the instrumentation field are under development at the present time.

The results achieved to date are very good considering the characteristics of the capacitor body material being used in the tuning elements. It is possible that the present difficulty of dielectric tuning—i.e., obtaining a large tuning ratio while maintaining a small temperature coefficient—may be considerably reduced through the development of new materials and manufacturing techniques.

[5] M. Apstein and H. H. Wieder, "Capacitor-modulated wide range FM system," *Electronics*, vol. 26, p. 190; October, 1953.

Note on the Design of Wide-Band Low-Noise Amplifiers*

D. WEIGHTON†

Summary—**The case is considered of a grounded cathode amplifier fed from a finite source impedance in which the requirements of minimum noise factor conflict with those of adequate bandwidth. Equalization, either by feedback or by the use of complementary networks, provides one means of dealing with this problem, and involves modification of some circuit parameters for minimum noise factor. Design criteria are developed for amplifiers based on this procedure, including equalization of both input and tube coupling circuits. Some measurements on an experimental amplifier are reported.**

Introduction

THE PRINCIPLES of design of low-noise amplifiers in which bandwidth is not a limiting factor are well-known as a result of the early work of a number of authors, notably North[1] and Herold.[2] In the case where the requirements of minimum noise factor conflict with those of adequate bandwidth there is, however, no clearly established procedure, although a number of devices to increase bandwidth have been discussed by Herold,[2] Lebenbaum[3] and others. In the narrow-band case, the input circuit of a grounded cathode amplifier is normally designed for minimum noise factor by choice of the optimum transformation of source impedance as described, for example, by Wallman[4] or Houlding.[5] The value of this optimum source impedance depends only on the circuit losses, the transit time damping and the equivalent noise resistance of the tube, and the maximum bandwidth available in these conditions is therefore fixed by the tube input and stray capacitances. If a greater bandwidth is required there are a number of alternatives open to the designer, but in most cases some sacrifice in noise factor is involved. A detailed discussion of these alternatives is to be found in Twiss and Beers.[6]

The use of feedback affords one of the most attractive solutions to this problem largely for incidental reasons. It allows some freedom in modifying the input impedance of the amplifier without appreciably degrading the noise factor and tends to minimize the effect of variations in source impedance. The effect of feedback on single frequency noise factor has been dealt with by Harris[7] and more recently by Beers,[8] who also discusses specific cases based on the work of Macnee. These authors show that, provided the feedback network does not itself significantly load the input circuit, the effect on single-frequency noise factor is small. The effect on the average noise factor over a wide pass band may, however, be large, since the spectral distribution of noise is changed, and a new value of optimum source impedance has to be found. The problem is similar to that of an amplifier in which the restricted bandwidth of the input circuit is corrected by an equalizing network inserted at some later stage. The latter arrangement has received some attention in relation to video amplifiers fed from a

* Original manuscript received by the IRE, January 7, 1955; revised manuscript received, June 17, 1955.

† Pye, Ltd. Radio Works, Cambridge, England.

[1] B. J. Thompson, D. O. North, and W. A. Harris, "Fluctuations in space-charge-limited currents at moderately high frequencies," *RCA Rev.*, vols. 4 and 5, pp. 269–285; January, 1940–July, 1941.

[2] E. W. Herold, "An analysis of the signal-to-noise ratio of ultra high frequency receivers," *RCA Rev.*, vol. 6, pp. 302–331, 1942.

[3] M. T. Lebenbaum, "Design factors in low noise figure input circuits," Proc. IRE, vol. 38, pp. 75–80; January, 1950.

[4] H. Wallman, A. B. Macnee, and C. P. Gadsden, "A low-noise amplifier," Proc. IRE, vol. 36, pp. 700–708; June, 1948.

[5] N. Houlding, "Noise factor of conventional V.H.F. amplifiers," *Wireless Eng.*, vol. 30, pp. 281–306; November, December, 1953.

[6] R. Q. Twiss and Y. Beers, "Vacuum Tube Amplifiers," M.I.T. Rad. Lab. Ser., No. 18, McGraw-Hill Book Co., Inc., New York, N. Y., ch. 13; 1948.

[7] W. A. Harris, "Fluctuations in vacuum-tube amplifiers and input systems," *RCA Rev.*, vol. 6, pp. 114–124; July, 1941.

[8] Y. Beers, "Microwave Receivers," M.I.T. Rad. Lab. Ser., No. 23, McGraw-Hill Book Co., New York, N. Y., p. 88; 1948.

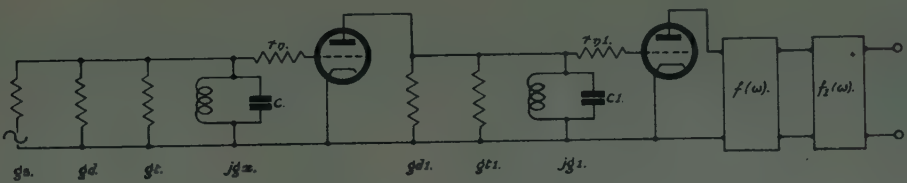


Fig. 1

constant current source (see, for example, Barco[9]) but as far as the author is aware it has not received quantitative treatment for the case of a source of constant available power.

The purpose of the present note is to develop design criteria and performance data for minimum-noise equalized amplifiers and to present the results in a form directly useful to the engineer. The discussion is limited to the grounded cathode circuit fed from a source of finite impedance, but applies equally to pentodes or triodes which may be either neutralized or cascode stabilized.

Input Circuit

The simplest arrangement of input circuit for a band-pass amplifier is shown in Fig. 1, where the total capacitance to ground is tuned by a shunt inductor. Writing g_s the source conductance, g_d the conductance representing circuit losses, g_t the transit time conductance, and g_x the total shunt susceptance, then the transfer function of the input circuit is of the form

$$g_s/(g_s + g_d + g_t + jg_x),$$

and the correcting network must therefore have the form

$$f(\omega) = (g_s + g_d + g_t) + jg_x, \tag{1}$$

if the amplifier is to be distortionless.

It is shown in Appendix A that the noise factor of the first stage of an amplifier which is followed by a correcting network of this form is given by

$$N = 1 + \left\{Kg_t + g_d + \frac{4}{3} r_n\pi^2B^2C^2\right\} \Big/ g_s + r_n(g_s + g_t + g_d)^2/g_s, \tag{2}$$

where K is the excess noise temperature assigned to the transit time loading, B is the over-all noise bandwidth of the amplifier in cycles per second and r_n is the equivalent noise resistance of the first tube. Eq. (2) will be recognized as the expression for single-frequency noise factor at band center with an additional term in the first bracket. It follows that the bandwidth limitation may be taken into account by supposing an additional noise source equal to that of a conductance $4/3\, r_n\pi^2B^2C^2$ to be in shunt with the input circuit.

In a wide-band amplifier g_d and g_t will generally be small compared with g_s, and the expression for noise factor may be approximated

$$N = 1 + \left(Kg_t + \frac{4}{3} r_n\pi^2B^2C^2\right) \Big/ g_s + r_ng_s. \tag{3}$$

In video amplifiers or intermediate frequency amplifiers in which the center frequency is sufficiently low, the bandwidth term predominates over the induced grid noise and the expression may be further approximated

$$N = 1 + 4r_n\pi^2B^2C^2/3g_s + r_ng_s, \tag{4}$$

from which the optimum source impedance is seen to be

$$r_s \text{ optimum} = \sqrt{3}/2\pi BC, \tag{5}$$

and the minimum noise factor,

$$N \text{ minimum} = 1 + 2.32r_n\pi BC. \tag{6}$$

It will be observed that under these conditions the optimum source impedance is independent of the equivalent noise resistance of the tube. Substituting from (5) into the transfer function of the input circuit, we obtain the result that the response falls by a factor of 2 at the edge of the band. The best noise factor is therefore obtained when the coupling of the source is adjusted to give an input circuit response which falls by 6 db at the edges of the desired band.

It is interesting to compare the performance of an amplifier designed on this basis with that of one in which the input circuit is damped by the source to a point where it is substantially flat over the desired band so that no compensation or feedback is necessary. For the sake of comparison substantially flat may be interpreted as dropping by 1 db at the limits of the over-all noise bandwidth. In this condition, $\pi BCr_s = .255$.

Rearranging (4),

$$N = 1 + r_n \cdot \pi BC \cdot \left\{\frac{g_s}{\pi BC} + \frac{4\pi BC}{3g_s}\right\},$$

$$N = 1 + 4.35\pi BCr_n.$$

Comparing this result with (6), it is seen that the incremental noise is improved by a factor of nearly 2 by the use of the correct source impedance combined with compensation. When the induced grid noise is not negligible as when the bandwidth is smaller or the center frequncy higher, the improvement effected in noise factor will be less.

Tube Coupling Circuits

The discussion has so far neglected the noise sources associated with the second and subsequent stages. The contributions to the total noise from these sources be-

[9] A. A. Barco, "An iconoscope preamplifier," *RCA Rev.*, vol. 4, pp. 89–107; July, 1939.

come of increasing importance as the bandwidth is made wider, and for minimum noise factor it is evidently desirable to achieve the greatest possible gain—bandwidth product for each stage. The noise relations in double-tuned circuits have been studied in this connection by Herold.[2] For optimum performance multiple circuit couplings tend to become critical of adjustment and difficult to handle in practice, and the case of an undamped circuit compensated by feedback is worthy of attention on the score of simplicity. The arrangement corresponds closely to the video amplifier described by Barco,[9] but the equations differ when the amplifier contains more than one circuit to be compensated. Integration of the noise over the band then involves cross products and the equations tend to become too complex to be useful to the designer. In most cases, however, the noise sources in the first two stages are the only ones of importance and the solution in this instance is derived in Appendix A.

Inspection of the expressions for noise factor given by (19) and (20) shows that the damping on the circuits g_d and g_{d1} should be as small as possible for the best noise factor. In the wide-band case an approximation may be made similar to that discussed above for the input circuit and (19) then simplifies to

$$N = 1 + r_s\left\{Kg_t + \frac{4}{3}\pi^2B^2C^2R_x\right\} + g_sR_y, \qquad (7)$$

where

$$R_x = r_n + \frac{1}{g^2}\left\{Kg_{t1} + \frac{12}{5}r_{n1}\pi^2B^2C_1^2\right\}$$

and

$$R_y = r_n + \frac{1}{g^2}\left\{Kg_{t1} + \frac{4}{3}r_{n1}\pi^2B^2C_1^2\right\}.$$

The contribution from the second tube appears in terms in both g_s and $1/g_s$ and will in general modify the optimum value of source impedance. However, the effect will be small unless the stage gain is very low, and the source impedance given by (5) may usually be used with negligible rise in noise factor. Substituting for g_s from (5) in (7), and neglecting the induced grid noise terms, we obtain

$$N = 1 + \frac{4\pi BC}{\sqrt{3}}\left\{r_n + \left(\frac{\pi BC_1}{.73g}\right)^2 r_{n1}\right\}. \qquad (8)$$

Comparing this with (6), it appears that for the purposes of calculating noise the effective voltage gain of the first stage is

$$\frac{.73g}{\pi BC_1}. \qquad (9)$$

When the input circuit is not equalized but is made sufficiently flat over the required frequency band by reducing the source impedance, then the term in $1/g_s$ in (7) will be small compared with the term in g_s. The effective voltage gain of the first stage is then obtained from the coefficient of r_{n1} in the term in g_s and will be

$$\frac{.87g}{\pi BC_1}. \qquad (10)$$

Comparing (9) and (10), it appears that the effective stage gain does not vary greatly for widely varying values of source impedance. The expressions are in a form permitting direct comparison with the performance of other types of coupling circuit,[10,11] for example, they compare favorably with the corresponding formula for a two-circuit coupler having unequal Qs.

Eq. (8) illustrates a feature of wide-band amplifiers which has not been emphasized in the literature. The criterion for low-noise factor in the first tube is a low value of the product r_nC, whereas in the second stage the contribution to the noise depends on $r_{n1}C_1^2$. This may have an important bearing on the selection of suitable tubes for these positions in a low-noise amplifier.

Matching to the Source

In narrow-band amplifiers the optimum source impedance is generally lower than the input impedance of the amplifier if the latter is largely determined by transit time damping. In the wide-band case the discrepancy will be greater, but the use of a feedback network to equalize the response of the input circuit suggests the possibility of also obtaining a match by this means. To satisfy the matching requirement and at the same time equalize with sufficient accuracy over the desired band it would theoretically be necessary to provide a rather complex feedback network. In the simplest case of a resistor connected back to the input circuit from a point in the amplifier where the gain is sufficiently high there are two parameters at the disposal of the designer, the feedback factor at band center and the group time delay round the loop. If the first of these is chosen to match the input impedance to the source, the second may be adjusted for a first-order equalization, but the frequency band over which such equalization is adequate will then be fixed. The result is, however, reasonably good for a source impedance chosen in accordance with (5), and it is shown in Appendix B that a maximally flat condition may then be obtained in which the response falls by ½ db at the limits of the design bandwidth. If the induced grid noise is significant so that the circuit to be equalized falls by less than 6 db at the band limits, then the matching condition may evidently be satisfied with an even better over-all response.

Practical Aspects

The construction of wide-band amplifiers in accordance with the above principles presents two major

[10] H. A. Wheeler, "Wide-band amplifiers for television," Proc. IRE, vol. 27, pp. 429–438; July, 1939.
[11] D. Weighton, "Performance of coupled and staggered circuits in wide band amplifiers," *Wireless Eng.*, vol. 21, pp. 468; October, 1944.

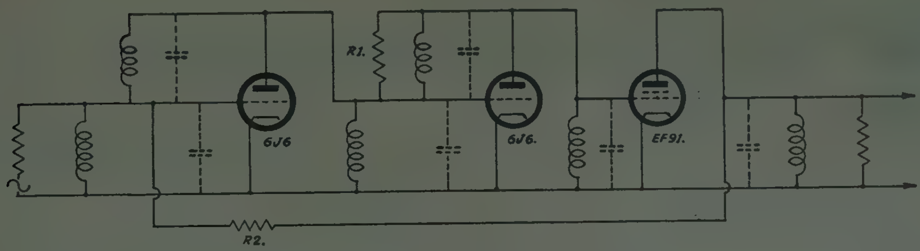


Fig. 2—Experimental amplifier showing ac connections only.

practical difficulties. The first is the familiar one of avoiding the effects of anode-grid capacitance in a grounded cathode triode, and the second the design of practical networks for compensation of the input and tube coupling circuits. An experimental amplifier has been constructed which offers one approach to the solution of these two problems. Compensation is by feedback rather than by complementary networks, since the response is then less critically dependent on the source impedance. In order to avoid appreciable damping on the input circuit by the feedback network itself, it is desirable that this should be taken from a point in the amplifier where the voltage gain is high. A similar remark applies to the circuit coupling the first and second tubes. However, since a compromise has to be effected and the input circuit is evidently the more important from the noise point of view, the arrangement in Fig. 2 (above) was tried and found to be reasonably satisfactory. Feedback for compensation of the second circuit is over only one stage via the resistor R_1 in shunt with a resonant circuit of which the plate-grid capacitance of the tube forms the major portion of the tuning capacitance. The arrangement can readily be adjusted for a maximally flat response without the addition of damping across the circuit. Feedback for the input circuit is over three stages allowing the resistor R_2 to be large compared with the source impedance. In one example for a bandwidth of 16 mc the optimum source inpedance was found to be about 500 ohms and R_2 was 18,000 ohms having no measurable effect on the single-frequency noise factor.

Inductive neutralization is employed throughout, since it affords the simplest circuit arrangement. The ease of neutralization in the first stage is greatly helped by the low voltage gain between grid and plate of the first tube. Since the feedback resistor R_1 is generally small compared with the plate impedance of the second tube, the input impedance of this stage at band center approximates to the reciprocal of the slope of the tube, and the voltage gain of the first stage is therefore one when the slopes of the two tubes are equal. In this respect the arrangement resembles the cascode. It differs in that the gain of the first stage rises on either side of band center. It is, however, sufficiently low at all times to ensure stability, and neutralization has been found to be very noncritical.

Experimental Results

No detailed experimental investigation has been undertaken, but the noise factor of one amplifier of the kind described above has been measured for a range of source impedance, adjusting the feedback in each case to maintain a flat response, and the results are shown in Fig. 3. The first two tubes were 6J6s with the two halves operated in parallel and the main parameters as follows:

Total shunt capacitance of input circuit	25pF.
Total shunt capacitance of coupling circuit	15pF.
Slope of the first tube	10,000 micromhos.
Equivalent noise resistance	300 ohms.
Transit time damping	20,000 ohms.
Total noise bandwidth	15 mc.
Band center frequency	35 mc.

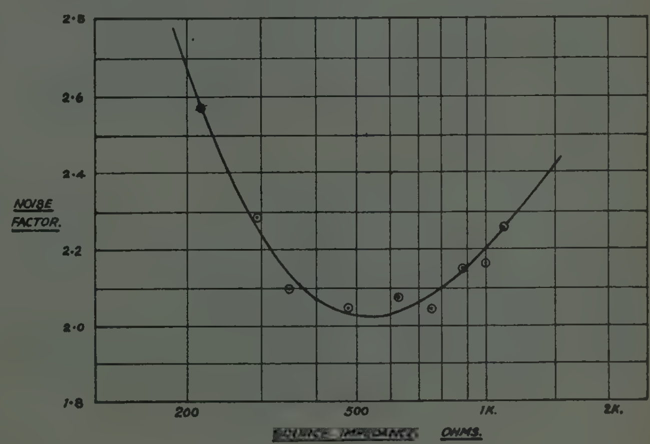


Fig. 3

Calculation of the anticipated performance illustrates the use of the equations and provides a check on the experimental results. The effective voltage gain of the first stage is first estimated from (9) and yields a figure of 10.3 times. The noise contribution from the second stage may therefore be neglected and the optimum source impedance derived from (3). Allowing a noise temperature ratio of 5 for the induced grid noise, the terms are

$$1/kg_t = 4{,}000 \text{ ohms,}$$

and

$$3/4r_n\pi^2B^2C^2 = 1{,}800 \text{ ohms,}$$

giving a total shunt resistance of 1,240 ohms. The optimum source impedance is therefore the geometric mean

of 1,240 ohms and the equivalent noise resistance 300 ohms, i.e., 610 ohms, and the corresponding noise factor

$$N = 1 + 2 \times \frac{300}{610}$$

$$N = 1.98.$$

The measured values show reasonably good agreement with these calculated figures.

Conclusions

Consideration of the noise relationships in grounded cathode wide-band amplifiers in which the input and tube coupling circuits are equalized by feedback reveals two points of interest to the design engineer.

1. The optimum source resistance may be selected by considering the usual narrow-band equation for noise factor with the addition of a hypothetical noise source equivalent to a conductance $4\, r_n \pi^2 B^2 C^2/3$ in shunt with the input. In the case of very large bandwidth or low mean frequency where this hypothetical source predominates over induced grid noise, the optimum condition is one in which the circuit falls by 6 db at the edges of the desired frequency band.
2. The equalized tube coupling circuit has an effective gain-bandwidth for noise calculations which compares favorably with that of multiple circuit networks.

There appears to be no major practical difficulty in constructing amplifiers of this kind which approximate closely to the theoretical limits at least for bandwidths up to about 20 mc.

Appendix A

Fig. 1 shows the essential features of an amplifier in which the input and first stage coupling circuits are followed by their appropriate correcting networks. From the transfer functions of the two circuits it is readily shown that the correcting networks have the form

$$f(\omega) = (g_s + g_d + g_t) + jg_x$$
$$f_1(\omega) = (g_{d1} + g_{t1}) + jg_{x1}. \qquad (11)$$

Neglecting any correlation between induced grid noise and shot noise, the equivalent mean-square noise voltage per cycle at the grid of the first tube is found to be

$$v^2 = 4kT(g_s + g_d + Kg_t)/\{(g_t + g_d + g_s)^2 + g_x^2\} + 4kTr_n, \qquad (12)$$

where KT is the noise temperature assigned to the transit time conductance, and r_n is the equivalent noise resistance of the tube.

The mean square noise voltage per cycle at the grid of the second tube due to the sources in the second circuit is given by

$$v_1^2 = 4kT(g_{d1} + Kg_{t1})/\{(g_{t1} + g_{d1})^2 + g_{x1}^2\} + 4kTr_{n1}. \qquad (13)$$

Now the mean noise power per cycle at the output of the amplifier is proportional to

$$g^2v^2\,|f(\omega)|^2 + v_1^2\,|f(\omega)|^2 \cdot |f_1(\omega)|^2, \qquad (14)$$

where g is the transconductance of the first tube. Substituting for v^2 and v_1^2 from (12) and (13), this expression becomes

$$4kT[g^2(g_s + g_d + Kg_t) + (r_ng^2 + g_{d1} + Kg_{t1}) \cdot |f(\omega)|^2 + r_{n1} \cdot |f(\omega)|^2 \cdot |f_1(\omega)|^2]. \qquad (15)$$

Substituting for $f(\omega)$ and $f_1(\omega)$ from (1) and (13) yields an expression containing terms in g_x^2, g_{x1}^2, $g_x^2g_{x1}^2$ and terms independent of frequency. To find the total noise this function must be integrated over the band accepted by the remainder of the amplifier, and the integrand should properly be modified in those regions where the over-all amplitude response is not flat. In practice it is usually sufficient to estimate the noise bandwidth and to integrate within these limits assuming a flat response.

If ω_0 is the angular frequency at the band center, then

$$g_x^2 = \omega_0^2C^2\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)^2,$$

and writing

$$p = (\omega_1 - \omega_2)/2\omega_0$$

where ω_1 and ω_2 define the limits of the noise bandwidth

$$\int_{\omega_1}^{\omega_2} g_x^2 \cdot d\omega = (\omega_2 - \omega_1)\omega_0^2C^2p^2(4 - p^2)/3(1 - p^2).$$

Provided p is less than about 0.25, a very good approximation is given by

$$\int_{\omega_1}^{\omega_2} g_x^2 \cdot d\omega = (\omega_2 - \omega_1) \cdot \frac{4}{3}\, p^2\omega_0^2C^2$$
$$= (\omega_2 - \omega_1) \cdot \frac{4}{3}\, \pi^2B^2C^2 \qquad (16)$$

where B is the noise bandwidth.

The remaining terms may be integrated in a similar manner, giving

$$\int_{\omega_1}^{\omega_2} g_{x1}^2 \cdot d\omega = (\omega_2 - \omega_1) \cdot \frac{4}{3}\, \pi^2B^2C_1^2 \qquad (17)$$

$$\int_{\omega_1}^{\omega_2} g_x^2 \cdot g_{x1}^2 \cdot d\omega = (\omega_2 - \omega_1) \cdot \frac{16}{5}\, \pi^4B^4C^2C_1^2. \qquad (18)$$

Using these relations in the integration of (15) and dividing by the noise originating in the source gives the noise factor:

$$N = 1 + r_s\left(g_d + K_{gt} + \frac{4}{3}\, \pi^2B^2C^2R_x\right) + r_s(g_s + g_d + g_t)^2R_y, \qquad (19)$$

where

$$R_x = r_n + \frac{1}{g^2}\left[g_{d1} + Kg_{t1} + r_{n1}\left\{(g_{d1} + g_{t1})^2 + \frac{12}{5}\pi^2 B^2 C_1^2\right\}\right]$$

$$R_y = r_n + \frac{1}{g^2}\left[g_{d1} + Kg_{t1} + r_{n1}\left\{(g_{d1} + g_{t1})^2 + \frac{4}{3}\pi^2 B^2 C_1^2\right\}\right]. \quad (20)$$

When the noise sources in the second stage are omitted,

$$R_x = R_y = r_n,$$

and inserting these values in (19) gives the noise factor of the first stage alone.

Appendix B

Fig. 4 shows an input circuit with current feedback via a resistor R; the amplifier having a gain of $-A$ and a time delay T. The analysis is carried out for the low-pass case but applies equally to a band-pass amplifier if T is the group delay.

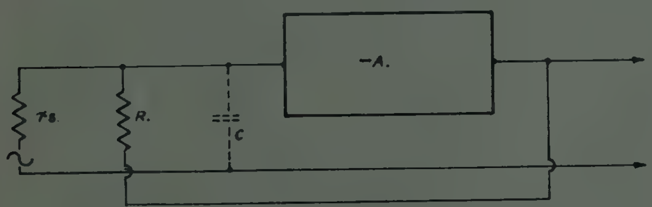


Fig. 4

The effective admittance due to feedback is of the form

$$\frac{1}{R}(1 + Ae^{-j\omega T}),$$

and the total admittance of the circuit is therefore

$$Y = \frac{1}{r_s} + j\omega C + \frac{1}{R}(1 + Ae^{-j\omega T}).$$

When the matching condition is satisfied and R is large compared with r_s,

$$Y = j\omega C + \frac{1}{r_s}(1 + e^{-j\omega T}).$$

Separating real and imaginary terms and taking the modulus gives

$$|Y|^2 r_s^2 = 2(1 + \cos\omega T) + \omega^2 C^2 r_s^2\left(1 - \frac{2\sin\omega T}{\omega C r_s}\right).$$

Expanding in powers of ω and equating the coefficient of ω^2 to zero, we obtain

$$T = (\sqrt{2} - 1)Cr_s.$$

This is the condition for first-order equalization. If r_s is chosen in accordance with criterion (5), then

$$\omega C r_s = \sqrt{3}$$

at the limits of the band and hence

$$\omega T = \sqrt{3}(\sqrt{2} - 1)$$
$$= .715 \quad \text{or} \quad 41°.$$

Substituting these values in the expression for transfer admittance, we obtain

$$|Y|^2 r_s^2 = 4.23.$$

When $\omega = 0$, $|Y|^2 r_s^2 = 4$, and the response at the edge of the band is therefore down in the ratio $\sqrt{4/4.23} = .945$, or about $\frac{1}{2}$ db.

Acknowledgment

The author is indebted to J. Blades for many useful discussions and to the Directors of Messrs. Pye Limited for permission to publish this paper.

Correction

H. A Haus and F. N. H. Robinson, authors of the paper, "The Minimum Noise Figure of Microwave Beam Amplifiers," which appeared on pages 981–991 of the August, 1955 issue of the Proceedings of the IRE, have brought the following corrections to the attention of the editors.

1. Eq. 3 should read

$$q = (q_1 e^{j\beta_p z} + q_2 e^{-j\beta_p z})e^{-j\beta_e z}.$$

2. Paragraph above Eq. 27 should read

The determinant of P is det $P = \pm 1$ and so (25) . . .

3. Delete this sentence, which follows Eq. 54:

The equality sign applies when $M_{14} = 0$, which will be the case if the amplifier presents a match to the output transmission line.

4. Eq. 61 should read

$$G(z) = Ge^{-2\beta_e C x_1 z}.$$

where $\beta_e C$ and x_1 have their usual meaning.[1]

5. The integral expression below Eq. 61 should read

$$\int_0^l e^{-2\beta_e C x_1 z}\lambda dz.$$

6. Immediately below this integral expression, change lower case c to capital C.

7. In Eq. 76 insert an equality sign between closed paren. and E_z.

8. Two lines below Eq. 84, $\epsilon_2 = \epsilon_2^*$ should read $\epsilon_1 = \epsilon_2^*$.

Wide-Range Electronic Tuning of Microwave Cavities*

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Summary—**Methods for electronically tuning microwave cavities using the principles of space-charge tuning and of spiral-beam electronic tuning in the presence of a low-pressure gas are described. The use of a low-pressure gas permits tuning over frequency ranges several times larger than those obtainable in vacuum.**

An *S*-band cavity was tuned over a frequency range from 3,280 to 4,350 mc, and from 3,280 to 2,540 mc, or ±30 per cent. An *X*-band cavity was tuned over a frequency range from 9,170 to 10,800 mc, or 18 per cent. These values are compared to measurements made in vacuum.

A semi-quantitative theory for electronic tuning in gas atmospheres is presented. Limitations of the method are given.

Introduction

Electronic tuning of microwave resonant elements is necessary for many applications in which frequency must be varied rapidly. For example, frequency-modulated oscillators may use electronically tuned microwave resonant cavities. Although earlier electronic-tuning techniques limited the tuning range of microwave cavities to a few per cent, the use of a low-pressure gas atmosphere has been found to permit the tuning of microwave cavities over frequency ranges several times larger than those obtainable when corresponding techniques are used in vacuum.

Two principal methods are used for the electronic tuning of microwave cavities in vacuum. The first method, spiral-beam tuning,[1,2] involves the projection of an electron beam into the resonant structure in a direction perpendicular to the oscillating electric field and parallel to a constant magnetic field, as in Fig. 1(a). The second method, space-charge tuning,[3–5] involves rotation of a cloud of electrons in a plane perpendicular to a constant magnetic field and in the same plane as the oscillating electric field, as in Fig. 1(b).

Both methods of tuning involve the interaction of rotating electrons with the radio-frequency electric field. Although this article describes the effects of this interaction in terms of changes in resonant frequency, the process is fundamentally an electronic means for changing capacitance. The various tuning methods described, therefore, are applicable not only to resonant elements but also to any other microwave network involving electric fields, i.e., capacitances. This paper discusses theoretical aspects and experimental results for spiral-beam tuning and space-charge tuning in gas atmospheres and compares these results to those obtained when the same tuning methods are used in vacuum.

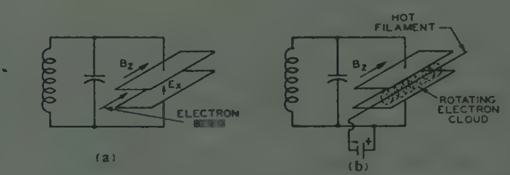


Fig. 1—(a) Schematic representation of resonant cavity using spiral-beam tuning. (b) Schematic representation of resonant cavity using space-charge tuning.

Theoretical Aspects

Spiral-Beam Tuning

In spiral-beam tuning, an electron beam is introduced into the concentrated electric-field region of a cavity, as shown schematically in Fig. 1(a). The magnetic field, which is parallel to the electron stream, causes the electrons to precess, i.e. to describe helical orbits. The electrons, therefore, exchange energy with the radio-frequency electric field, E_x, inducing a reactive current in the cavity walls and thereby having a tuning effect on the cavity. The period of rotation of the electrons is determined by the magnetic field in accordance with the formula $\omega_c = (|e|/m)B_z$, where ω_c is the angular frequency of rotation in radians per second, $|e|$ and m are the charge and mass, respectively, of an electron, and B_z is the magnetic-field density in webers per square meter. Depending on the value of ω_c and the combination of angular frequency, ω, of the rf field, geometry, and electron entrance velocity, v_0, the electron beam behaves as an electronically variable admittance, with or without a dissipative component. This admittance is either inductive or capacitive, i.e., can either raise or lower the resonant frequency of the circuit electronically. This relation has been analyzed by Smith and Shulman[1] and by Baños and Saxon.[2]

Smith and Shulman begin their analysis with the electronic equations of motion:

$$m\ddot{x} = -E_0|e|e^{i\omega t} - B_z|e|\dot{y} \tag{1}$$

$$m\ddot{y} = +B_z|e|\dot{x} \tag{2}$$

$$m\ddot{z} = 0 \tag{3}$$

where x, y, and z are the three Cartesian co-ordinates, t is time in seconds, $E_0 e^{i\omega t}$ represents the electric-field intensity varying at an angular radio frequency $\omega(=2\pi f)$, B_z is the constant magnetic-field density, and the dots above the symbols x, y, and z, denote first and second derivatives with respect to time. In this paper, a distinction is made between ω, the resonant angular frequency of the tuned cavity, and ω_0, the resonant angular frequency of the untuned cavity, such that $\omega = \omega_0 + \Delta\omega$

* Original manuscript received by the IRE, February 2, 1955; revised manuscript received, June 6, 1955.
† Radio Corporation of America, Tube Division, Harrison, N. J.
[1] L. P. Smith and C. Shulman, "Frequency modulation and control by electron beams," Proc. IRE, vol. 35, pp. 644–657; July, 1947.
[2] A. Baños, Jr., and D. S. Saxon, "An Electronic Modulator for CW Magnetrons," M.I.T. Rad. Lab. Rep. 748; June 26, 1945.
[3] J. P. Blewett and S. Ramo, "High frequency behavior of a space charge rotating in a magnetic field," *Phys. Rev.*, vol. 57, pp. 635–641; 1940; and *Jour. Appl. Phys.* vol. 12, pp. 856–859; 1941.
[4] W. E., Lamb, Jr., and M. Phillips, "Space-charge frequency dependence of a magnetron cavity," *Jour. Appl. Phys.*, vol. 18, pp. 230–238; February, 1947.
[5] H. W. Welch, Jr., G. R. Black, G. R. Brewer, and G. Hok, Final Report, "Theoretical Study, Design and Construction of CW Magnetrons for Frequency Modulation," Contract W-36-039-SC-32245, Univ. of Michigan, Engrg. Res. Inst., May 27, 1949.

For the initial conditions, Smith and Shulman assumed that when $t=t_0$, $x=z=0$, $\dot{x}=v_x=\dot{y}=0$, and $\dot{z}=v_0$. Thus, they obtained for the x-component of velocity:

$$v_x = -E_0 \frac{|e|}{m} \frac{i\omega}{\omega^2-\omega_c^2}\left\{\left(\frac{\omega_c+\omega}{2\omega_c}\right)e^{i(\omega_c-\omega)(t-t_0)} + \left(\frac{\omega_c-\omega}{2\omega_c}\right)e^{-i(\omega+\omega_c)(t-t_0)} - 1\right\}e^{i\omega t}. \quad (4)$$

If the frequency change is restricted so that the fraction $(\omega_c-\omega)/\omega$ is considerably less than unity, an expression for electron admittance as a function of the transit angle, θ, is obtained, as shown in Fig. 2. The

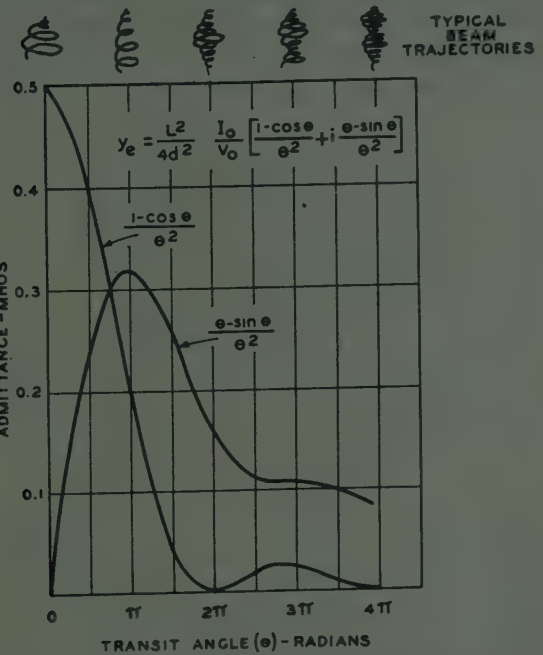


Fig. 2—Electronic admittance as a function of transit angle, θ.

transit angle, θ, in radians, is equal to $(\omega_c-\omega_0)\tau$, where τ is the electron transit time in seconds. Typical electron trajectories appear in Fig. 2 for several values of θ, when cavity is excited in the TE_{111} mode. When the electron enters the rf-field region, it is accelerated by the rf field toward one ridge (Fig. 7) and then begins to describe a circular path due to influence of the dc magnetic field. The length of time required for an electron to describe a complete circle is independent of the rf field, and is a function of the magnetic field only. Angular velocity of the electron, ω_c, is a constant equal to $B_z|e|/m$. If the angular frequency, ω, of the rf field is somewhat greater than ω_c, the time of one complete revolution is somewhat more than the time of one rf cycle so that the electron will lag somewhat behind the field. At the end of each succeeding revolution, the electron has gained more energy and has increased its radius of rotation, but it continues to fall further behind the rf field until its rotation is in quadrature with the field. Beyond this point, as the electron continues to rotate, it begins to give up some of its rotational energy to the rf field, and its radius of rotation decreases.

If the electron leaves the interaction space when its radius of rotation is zero ($\theta=2\pi n$; $n=1, 2, 3, \cdots$), the electron beam can be represented by a pure susceptance; this condition represents frequency modulation. If the electron leaves the interaction space at any other point ($\theta\neq 2\pi n$), the beam has rotational energy (which is dissipated as heat) and can be represented by a susceptance plus a dissipative component; this condition represents a combination of amplitude and frequency modulation. In the special case where $\omega=\omega_c$, the electron rotation is in phase with the rf field; as a result, the radius of rotation will increase without limit until the electron strikes the ridge; this condition represents amplitude modulation.

Even though the susceptance decreases with increasing transit angle, θ, (for $\theta>\pi$) as shown in Fig. 2, it is not inversely proportional to transit time, τ, because τ is a variable dependent on the electron entrance velocity v_0. Since

$$V_0 = \frac{1}{2}\frac{m}{|e|}\frac{L^2}{\tau^2},$$

by substituting for V_0 in the expression for electronic admittance shown in Fig. 2, the curves shown in Fig. 3, in a three-dimensional plot, are obtained. These curves indicate that for a given value of magnetic field the electronic susceptance increases continuously with transit time.

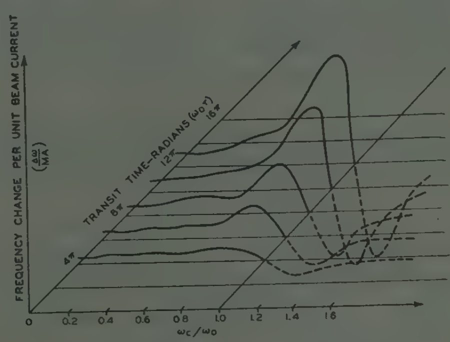


Fig. 3—Three-dimensional plot of theoretical frequency shift in spiral-beam tuning as a function of magnetic-field parameter, ω_c/ω_0, and transit time, $\omega_0\tau$.

In the experimental work described in this article, the experimental frequency shifts were large enough so that the restriction of frequency change $\omega_c-\omega\ll\omega$ was no longer valid. Therefore, new expressions for electronic admittance are shown in (7) and (8).

Spiral-Beam Tuning in a Gas Atmosphere

The electron is shot into the tuning cavity with a kinetic energy v_0, which is several times the ionization potential of the gas (15.7 volts for argon $A+$). After traveling a mean free path which can be calculated statistically, the electron experiences inelastic collisions with gas molecules, thereby producing ionization electrons. These then begin to spiral in phase with the rf field (as the primary electrons did when they entered the cavity), and thereby contribute to the electronic susceptance. Because the primary electrons are slowed

down by the collisions and the ionization electrons have very low drift velocity, the time during which the electrons interact with the rf field is substantially increased. Therefore, we can expect an increase in electronic reactance, over that obtainable with spiral-beam tuning in vacuum, due to two factors: (1) more electrons, and (2) increased interaction time, i.e., transit time.

More Electrons. An electron having an entrance velocity of 100 volts can produce as many as six ionization electrons in argon. The probability of an electron colliding with a gas molecule in a given distance in the z-direction is substantially increased by the magnetic field because of the spiral-beam helical motion of electrons in the (transverse) xy plane. Hence, a given mean free path may be divided by a factor, K, to obtain an equivalent mean free path which allows for increased path length due to the transverse helical electron motion. The value of the factor K is derived as follows:

The length, S, of the helical path of the electron in spiral-beam tuning for one spiral may be expressed as

$$S=2\int_0^{1/2(f_c-f)} \cdot\sqrt{\left(\frac{2|e|E_0\omega}{m(\omega_c^2-\omega^2)}\right)^2 \sin^2\frac{1}{2}(\omega_c-\omega)t+v_0^2}\quad dt. \tag{5}$$

If the electron moves in the (transverse) xy-plane in a spiral path without motion in the (axial) z-direction, i.e., v_0 is taken to be zero, (5) can be easily integrated and the length of the path between adjacent points of minimum radius obtained:

$$S = 8\frac{|e|}{m}\frac{E_0}{\omega^2}\left[\frac{1}{\left(\frac{\omega_c}{\omega}\right)^2-1}\right]\left[\frac{1}{\left(\frac{\omega_c}{\omega}\right)-1}\right]. \tag{6}$$

The axial distance, l, for one complete helical path, is given by

$$l = v_0 t = \frac{v_0}{f_c - f}.$$

The value of the factor K is approximately equal to the sum of the two path lengths (added linearly) divided by l:

$$K \approx \frac{S+l}{l}.$$

The equivalent mean free path, therefore, is a function of both the rf electric field and the electron-beam velocity.

Fig. 4 shows K as a function of beam voltage, V_0, for an S-band cavity fed from a local oscillator having an output of 50 milliwatts and for an X-band cavity fed from a 200-watt source. For the X-band cavity, K is about 10 for a 10-volt electron beam; therefore, for this case the helical electron motion increases the probability of ionizing collisions approximately tenfold.

Increased Interaction Time. The decrease in axial drift velocity ...

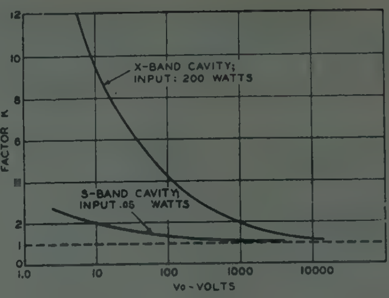


Fig. 4—Factor K as a function of beam voltage, V_0.

spiraling electron cloud due to collisions and the low axial drift velocity of the ionization electrons cause an increase in interaction time. The equation given by Smith and Shulman for the electronic admittance of the spiraling electron cloud may be solved in terms of transit time, τ, for the electronic conductance

$$G_e = \frac{|e|}{m}\frac{I_0}{2d^2}\left\{\frac{1}{(\omega_c-\omega)^2}\left[1-\cos(\omega_c-\omega)\tau\right] + \frac{1}{(\omega_c+\omega)^2}\left[1-\cos(\omega_c+\omega)\tau\right]\right\}, \tag{7}$$

and for the electronic susceptance

$$B_e = \frac{|e|}{m}\frac{I_0}{2d^2}\left\{\frac{1}{(\omega_c-\omega)^2}\left[(\omega_c-\omega)\tau-\sin(\omega_c-\omega)\tau\right] - \frac{1}{(\omega_c+\omega)^2}\left[(\omega_c+\omega)\tau-\sin(\omega_c+\omega)\tau\right]\right\}. \tag{8}$$

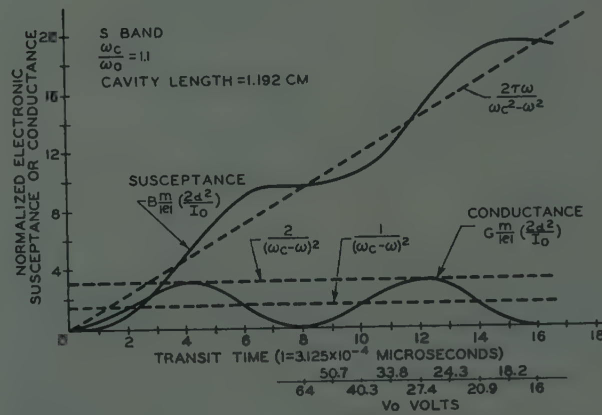


Fig. 5—Electronic admittance of spiral beam as a function of transit time.

Eqs. (7) and (8) are plotted in Fig. 5 in terms of normalized expressions

$$G\frac{m}{|e|}\frac{2d^2}{I_0}$$

and

$$B\frac{m}{|e|}\frac{2d^2}{I_0}$$

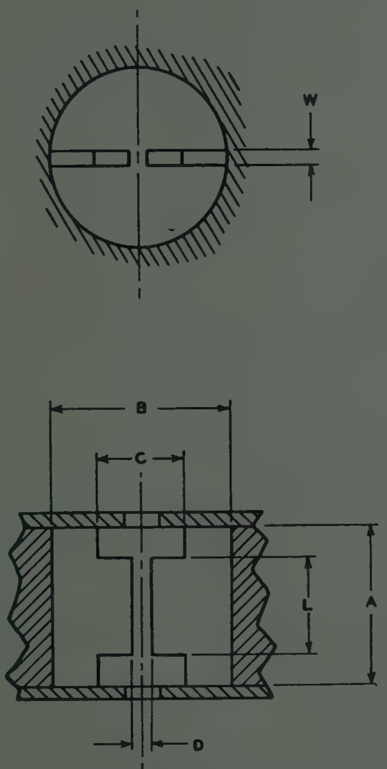
for a frequency of 4,000 megacycles per second and a ratio of ω_c/ω_0 equal to 1.1. The second term of (4), which is neglected by Smith and Shulman, is not neglected in

As shown in Fig. 5, the susceptance increases in an essentially linear manner with transit time while the conductance oscillates about a mean value. Therefore, even if a *slow-moving* electron does leave the interaction space with some rotational energy, the dissipative effect (AM) is very small in comparison to the reactive effect (FM). Thus, for a given frequency change, it becomes less and less important to maintain the transit angle, θ, equal to $2\pi n$ as the value of τ increases.

Experimental Results

Test Cavities

All tests reported in this article were made in ridge-waveguide resonant cavities having a length equal to one-half the guide wavelength. Fig. 6 shows the important dimensions of the cavities used for the S- and X-band tests.



CAVITY DIMENSIONS IN INCHES

	S-BAND	X-BAND
A	1.300	.660
B	1.500	.835
C	.700	
D	.130	.100
L	.800	.660
W	.100	200

Fig. 6—Ridge-waveguide resonant cavity used for S- and X-band tests.

In the S-band cavity, two coaxial feed lines which are coupled into the cavity by means of inductive loops allow both transmission-type and reaction-type measurements. The X-band spiral-beam cavity is shown in Fig. 7. The gun consists of a tetrode having 0.200-inch by 0.050-inch rectangular apertures. The indirectly heated oxide cathode operates at a temperature of 830 degrees C. and a heater voltage and current of 6.3 volts and 2 amperes, respectively.

The X-band cavity, which is coupled to 1-inch by $\frac{1}{2}$-inch waveguide through slots, has the following electrical values in the absence of the electron beam: $f_0=9,600$ mc; $Q_0=1,760$; $Q_L=380$. Vacuum-tight ceramic windows and mica windows are used in the waveguide to separate the cavity from the atmosphere.

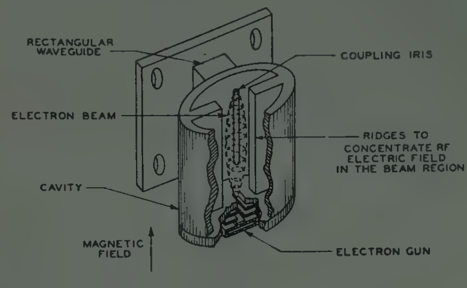


Fig. 7—X-band cavity using spiral-beam tuning.

Test Arrangements

Fig. 8 illustrates the setup used at low power levels to measure frequency changes in the X-band cavity by the reaction method. For the work at S-band, a simpler, transmission-type arrangement was used. An arrangement similar to that shown in Fig. 9 was used for tuning measurements at high power levels and for gas-breakdown tests.

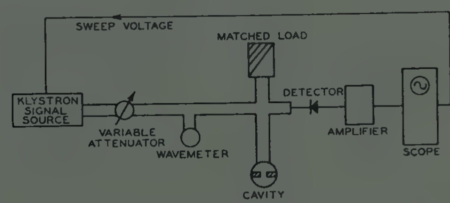


Fig. 8—Test setup used to measure frequency changes at low power levels in the X-band cavity.

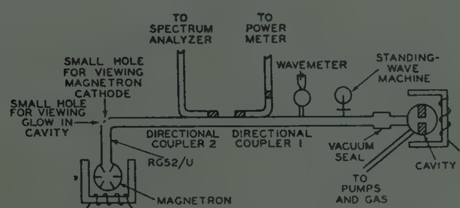


Fig. 9—Test setup used at high power levels and for gas-breakdown tests.

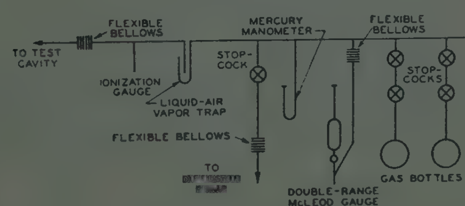


Fig. 10—Glass system used for exhaust and for gas-pressure measurements.

A special glass system, shown schematically in Fig. 10, provided the facilities required for exhaust, for filling the test cavities with gas, and for measuring the gas pressure over a wide range. The double-range McLeod gauge is designed and constructed to measure pressure continuously from 5 mm to about 0.2 microns of Hg.

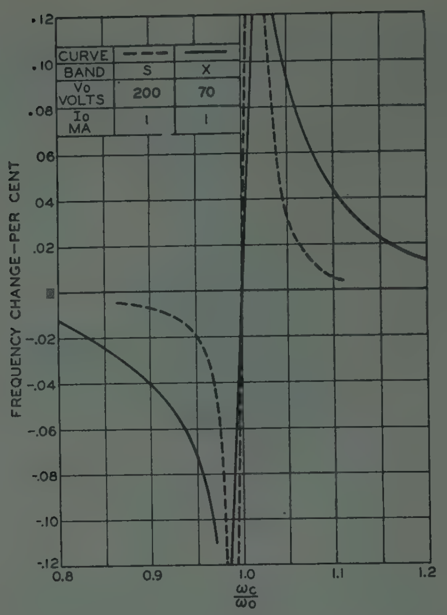


Fig. 11—Frequency change as a function of magnetic-field parameter for spiral-beam tuning in vacuum.

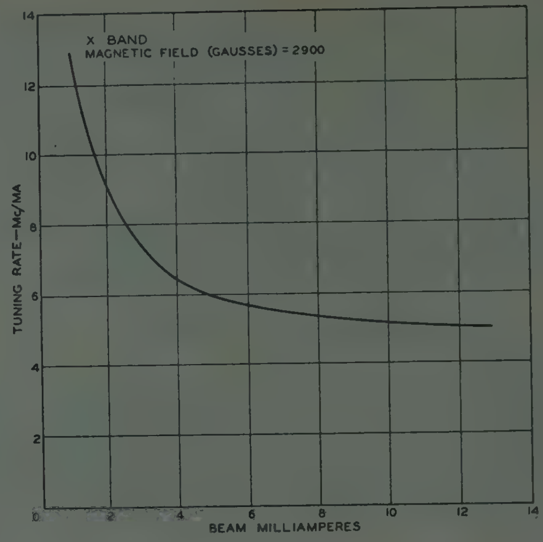


Fig. 12—Tuning rate as a function of beam current for spiral-beam tuning in vacuum.

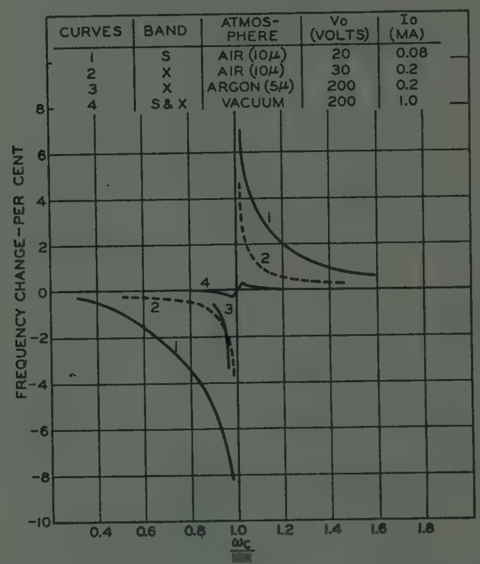


Fig. 13—Frequency change as a function of magnetic-field density for spiral-beam tuning in gas atmospheres, and comparison to results with vacuum.

Spiral-Beam Tuning in Vacuum

Fig. 11 shows the frequency change as a function of magnetic-field density with electron-beam current held constant for both *S*-band and *X*-band cavities using spiral-beam tuning in vacuum. At *S*-band (3,250 mc), a maximum frequency change of ± 30 mc (± 0.9 per cent) was obtained with a beam current of 2 milliamperes. (The plus and minus signs refer to settings of magnetic field for $\omega_c/\omega_0 < 1$ and >1, respectively.) The tuning rate was 15 mc/ma at 2 ma.

For a special test in the *S*-band cavity, a second electron gun was installed opposite the gun shown in Fig. 7, and two electron beams were shot into the cavity from opposite directions. As expected, the frequency deviation obtained was the same when each of the two guns was operated at one-half the current of the single gun used in the initial test. A maximum frequency change of ± 175 mc (± 5.5 per cent) was obtained with a total beam current of 20 milliamperes. This frequency change represents a tuning rate of 8.8 mc/ma at 20 ma.

At *X*-band (10,000 mc), a maximum frequency change of ± 500 mc (± 5 per cent) was obtained with a beam current of 14 ma. The tuning rate was 35 mc/ma at 14 ma. Although higher values of deviation can be obtained by the use of higher beam currents, the tuning rate decreases with increasing beam current, as shown in Fig. 12.

Spiral-Beam Tuning in Gas Atmosphere

Tests made in *S*-band and *X*-band cavities filled with low-pressure air showed that substantially larger frequency deviations could be obtained in low-pressure air than in vacuum. Fig. 13 shows the frequency deviation as a function of magnetic field for extremely low values of beam current and a probable air pressure of about 10 microns of Hg (not measured). In general, the curves obtained in air are similar to those obtained in vacuum, although there is some beam loading at magnetic-field values at which no loading would exist in vacuum.

Fig. 14 shows curves of wavelength versus beam current obtained in a spiral-beam *S*-band cavity filled with argon at various pressures. Two values of magnetic field were used: $\omega_c/\omega_0 = 0.86$ for the two upper curves, and $\omega_c/\omega_0 = 1/0.86 = 1.16$ for the lower curves. At a pressure of argon equal to 46 microns Hg and with a gun-cathode current of less than 1 milliampere, the frequency change was almost -30 per cent when the magnetic field was set for $\omega_c/\omega_0 = 0.86$ and $+30$ per cent for $\omega_c/\omega_0 = 1.16$. The cavity was actually tuned from 2,540 to 4,350 megacylces per second. It should be noted that the curves obtained for values of ω_c/ω_0 equal to 0.86 and

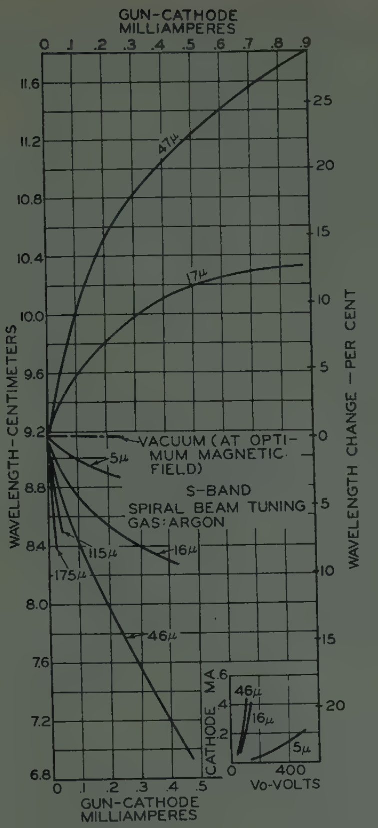


Fig. 14—Frequency as a function of gun-cathode current for a spiral-beam *S*-band cavity filled with argon at various pressures for two fixed values of magnetic field.

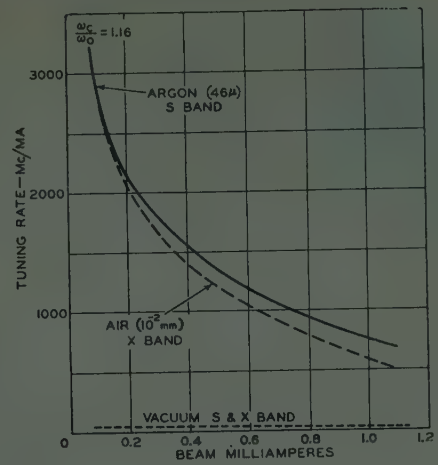


Fig. 15—Tuning rate as a function of beam current for spiral-beam tuning in gas atmospheres.

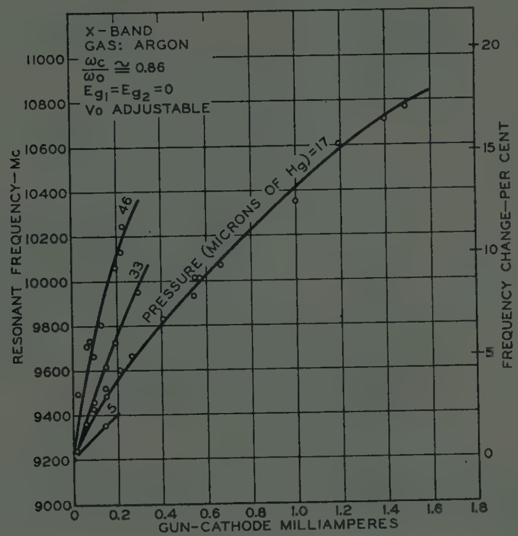


Fig. 16—Frequency change as a function of gun-cathode current for different values of pressure and fixed magnetic field.

1.16 are symmetrical. Although no measurements were made of the dissipative component, qualitatively it was found to be approximately proportional to pressure and low at pressures up to 46 microns of Hg. The tuning rate at a pressure of 46 microns of Hg is 760 mc/ma at 1 ma and 2,850 mc/ma at 0.1 ma. The tuning rate is shown in Fig. 15 as a function of beam current. A similar curve for spiral-beam tuning in vacuum is also shown for comparison.

The frequency change in an *X*-band cavity filled with argon is shown in curve 3 of Fig. 13 as a function of magnetic field for a constant beam current. Fig. 16 shows the frequency change at *X*-band as a function of the beam current for different values of pressure and fixed magnetic field ($\omega_c/\omega_0 = 0.86$). In all tests in which the cavities were filled with low-pressure air or argon, the limit on the frequency range obtained was determined by the test equipment rather than the test cavity. Therefore, frequency changes larger than those measured are obtainable.

The data shown in Figs. 11 through 16 were measured using a low-power tunable klystron oscillator having a power output of only a few hundred milliwatts as a signal source.

Comparison of Theory and Experiment for Spiral-Beam Tuning in Gas

In order to compare theory and experiment, we need to derive an expression for frequency change that is usable with the large values of frequency deviation observed with spiral-beam tuning in gas atmospheres. Eq. (8) for electronic susceptance, can be approximated by the first term, so that

$$B_e \cong \frac{|e|}{m} \frac{I_0}{2d^2} \left(\frac{\tau}{\omega_c - \omega} \right). \tag{9}$$

The resonant frequency may be determined by the application of the condition $\Sigma B = B_e + B_c = 0$, and by the use of the expression $B_c \cong 2C_0 \Delta\omega$ for circuit susceptance. Then, since $\omega = \omega_0 + \Delta\omega$,

$$\Delta\omega = \frac{(\omega_c - \omega_0) \pm \sqrt{(\omega_c - \omega_0)^2 + \frac{|e|}{m} \frac{I_0}{d^2} \frac{\tau}{C_0}}}{2}. \tag{10}$$

The positive sign may be eliminated by substitution of the condition $\Delta\omega = 0$ when $\tau = 0$, so that the fractional frequency change becomes

$$\frac{\Delta\omega}{\omega_0} = \frac{1-\omega_c/\omega_0}{2} \cdot \left[\sqrt{1+\frac{1}{(1-\omega_c/\omega_0)^2}\frac{|e|}{m}\frac{I_0\tau}{\omega_0^2 d^2 C_0}}-1\right]. \qquad (11)$$

This expression is usable for large values of $\Delta\omega$, and reduces, for the case of small frequency deviations, to

$$\frac{\Delta\omega}{\omega_0} = \frac{1}{1-\omega_c/\omega_0}\frac{|e|}{m}\frac{I_0\tau}{4\omega_0^2 d^2 C_0}. \qquad (12)$$

Curves of frequency change versus transit time are shown in Fig. 17 for values of ω_c/ω_0 equal to 1.1 and 0.9. The dashed lines in Fig. 17 represent (12). The error introduced by this approximation is appreciable for the physical conditions chosen even at a deviation of 10 per cent.

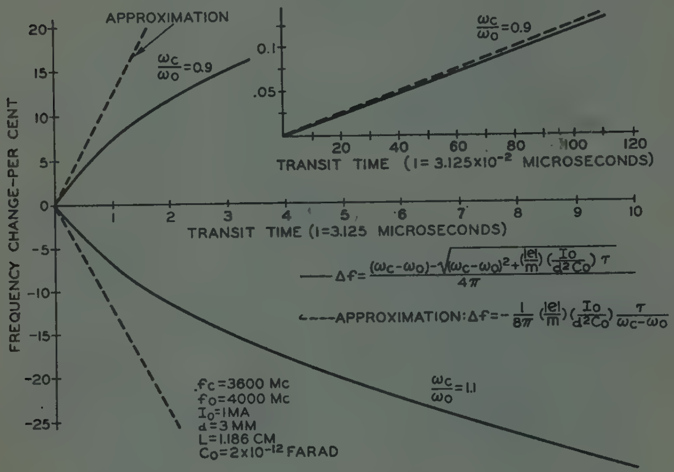


Fig. 17—Theoretically calculated frequency change as a function of transit time for spiral-beam tuning. Dashed lines represent the approximation $\Delta\omega \ll \omega$.

When physical constants and dimensions for the S-band cavity are substituted in (11), and a value of 0.86 is used for ω_c/ω_0,

$$\frac{\Delta\omega}{\omega_0} = 0.07\left[\sqrt{1+333\frac{I_0}{\sqrt{V_0}}}-1\right]. \qquad (13)$$

For a typical value of $I_0/\sqrt{V_0}$ equal to 10×10^{-6} (for example, $I_0=100$ microamperes; $V_0=100$ volts), the calculated frequency change, $\Delta\omega/\omega_0$ is 0.0117 per cent. This value is representative of the frequency deviation obtainable in vacuum for this current and voltage, but does not agree with the measured $\Delta\omega$ for gas-filled cavities because the measured value of cavity current, I_0, substituted in (13) is essentially a measure of primary electron flow alone. Eq. (13), therefore, should be rewritten with an additional term as follows:

$$\frac{\Delta\omega}{\omega_0} = 0.07\left[\sqrt{1+333\left(\frac{I_0}{\sqrt{V_0}}+\frac{nI_0}{\sqrt{V_2}}\right)}-1\right], \qquad (14)$$

where V_0 is used as a measure of the average transit time of the primary electrons, n is the number of ionization electrons produced per primary electron (and is a function of V_0) and V_2 is used as a measure of the average transit time of the ionization electrons.

Because the ionization electrons have very low initial velocities and there are more of them than primary electrons, the term in (14) involving V_0 is much smaller than the term involving V_2 and can be neglected.

For an S-band spiral-beam cavity filled with argon at a pressure of 46 microns of Hg, the frequency change for a beam current, I_0, of 100 microamperes is calculated as follows: The curves in Fig. 14 indicate that the measured value of beam voltage, V_0, for a beam current of 100 microamperes and a pressure of 46 microns of Hg is approximately 55 volts. Because about 10 ions, /cm/mm Hg are created in argon for each primary electron having a velocity from 50 to 100 volts,[6] at a pressure of 46 microns of Hg approximately one ion is created for two centimeters traversed by a primary electron. This value must be multiplied by the factor K to allow for the decrease in mean free path due to the spiral motion of the primary electron. For a value of K equal to 2 (from Fig. 4) and a cavity length, L, of about 2 centimeters, the number of created ionization electrons, n, for each primary electron is approximately 2. If it is assumed that the ionization electrons are at room temperature, their thermal velocity, $V_2=KT/2|e|$ is equal to 0.013 volt. By substitution in (12) a frequency deviation of 1.6 per cent is obtained. Although this value does not agree with the measured value of 8 per cent, it comes much closer than the value of 0.0117 per cent originally calculated.

Space-Charge Tuning

Phase shifts in a coaxial line using a rotating space charge in vacuum have been measured by Blewett and Ramo.[3] The problem of a rotating space charge in a nonoscillating magnetron has been investigated by Lamb and Phillips[4] and by Welch, *et al.*[5]

In the method of space-charge tuning shown in Fig. 1(b), a thin heated filament is placed between the two ridges parallel to a constant magnetic field, and a dc voltage is applied between the filament (which serves as the cathode) and the ridges (which serve as the anode). Space-charge tuning can be considered as closely analogous to spiral-beam tuning, particularly when the cathode is extremely small. In space-charge tuning, however, two modifications are: (1) $v_0=0$, and (2) a dc field is superimposed on the interaction space.

The dissipative component can be expected to be greater for space-charge tuning than for spiral beam tuning because the dc field attracts the electrons to the anode and sweeps them out of the interaction space while they possess rotational energy originating from the rf field. One therefore would not expect that there

[6] A. Von Engel and M. Steenbeck, "Elektrische Gasentladungen," J. Springer, Berlin, Germany; 1932.

exists a condition for zero dissipative component (pure FM) as in spiral beam tuning.

The test cavity used in the measurement of space-charge tuning was similar to the spiral-beam cavity shown in Fig. 7 except that a fine tungsten wire, mounted midway between the ridges, is substituted for the electron gun. The external dc magnetic field is parallel to the filament. All measurements were made at frequencies near 4,000 megacycles.

Curves 1 and 2 in Fig. 18 show the change in frequency obtained in a space-charge cavity in vacuum as a function of anode current for two values of ω_c/ω_0. A maximum deviation of +130 mc (+4 per cent) at 3,280 mc was measured. The dissipative component was found to increase rapidly in proportion to frequency deviation. Curves 3 and 4 show frequency change for space-charge tuning in argon as a function of anode current for two values of ω_c/ω_0. In argon frequency deviations of the order of +12 per cent were obtained. Curves 3 and 3a in Fig. 18 taken with different values of heater power show that space-charge tuning is very sensitive to filament temperature.

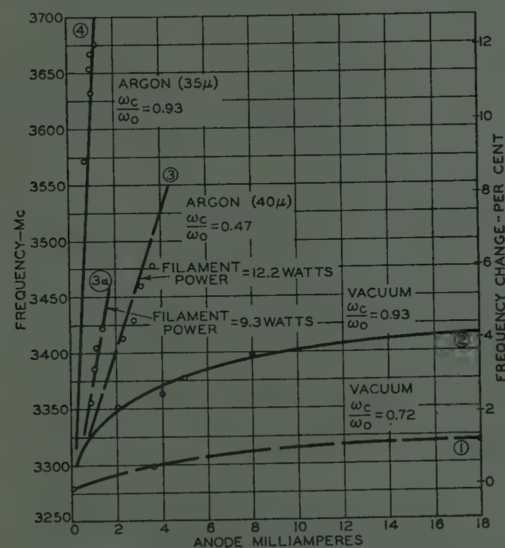


Fig. 18—Frequency change as a function of anode current for space-charge tuning for vacuum and gas atmosphere.

In space-charge tuning, rf absorption increases rapidly in proportion to frequency deviation. Electrons absorb energy from the rf and dc electric fields, spiral in increasingly larger orbits, and finally strike the anode, thereby dissipating energy in the anode in the form of heat. This loading and the sensitivity to filament temperature mentioned above are serious disadvantages of the space-charge tuning method both in vacuum and in gas. The use of an interaction space which is free of electrostatic fields and the separation of the electron-emitting means from the microwave portion of the device combine to make the spiral-beam tuning technique superior to the space-charge tuning method for both vacuum and gas atmosphere.

Limitations

Any tuning method which employs a gas atmosphere is subject to limitation by gas *breakdown*.[7] The breakdown point is that point at which cumulative ionization is initiated by the rf energy stored in the cavity. Beyond this point, the dc voltages have no control over the cavity frequency. The operating rf power level, therefore, must be kept below the power level necessary for initiating breakdown.

The presence of the magnetic field enhances breakdown. As ω_c approaches ω_0, the angular momentum of the electron continues to increase and the rotational energy becomes greater than the ionization potential for the gas used. For example, if the electronic maximum rotational energy, V_ω, is given by

$$V_\omega = \frac{1}{2}\frac{m}{|e|}\omega_c^2 r^2_{\max}, \tag{15}$$

and the maximum radius, $r_{\max}$ of the spiral beam is

$$r_{\max} = 2\frac{E_0|e|}{m}\frac{1}{\omega^2}\left\{\frac{1}{\left(\frac{\omega_c}{\omega_0}\right)^2 - 1}\right\} \tag{16}$$

then for the conditions $f=3{,}000$ mc, $\omega_c/\omega_0=1.1$, and $E_0=300$ volts/centimeter (corresponding to 50 milliwatts input power to the cavity from the signal generator), a value of V_ω equal to 20.6 volts is obtained. This value exceeds the ionization potential of argon. It must be concluded from this and from other tests that gas tuning can be used only in applications where the power level is of the order of one watt or less.

The magnitude of the *conductance component* was not measured during the frequency-deviation measurements. However, it has been observed qualitatively that the loading is not severe for spiral-beam tuning at gas pressures up to about 47 microns of Hg.

Gas pressure can be expected to decrease with the life of the tube. This decrease in pressure can be reduced to some extent by inclusion of a large glass bulb in the tube to serve as a gas reservoir, or by the use of hydrogen reservoir or a metallic vapor.

Modulation due to *plasma oscillations* or *noise* was not investigated.

Conclusions

The energy interaction of electrons and radio-frequency electric fields in resonant structures, such as microwave cavities, provides a means for producing electronically-controlled amplitude modulation and/or frequency modulation (tuning).[8]

Spiral-beam tuning in vacuum has been successfully applied to obtain cavity-tuned magnetron frequency

[7] B. Lax, W. P. Allis, and S. C. Brown, "Effect of magnetic field on the breakdown of gases at microwave frequencies," *Jour. Appl. Phys.*, vol. 21, pp. 1297–1304; December, 1950.

[8] F. Arams, "Microwave applications of gas discharges," *Electronics*, vol. 27, pp. 168–172; November, 1954.

changes of the order of 1 per cent.[9,10] No indication has been observed that this value can be substantially exceeded in vacuum in self-excited oscillators where a substantial fraction (such as 50 per cent) of the total energy is stored in the tuning cavity. However, it can be expected that more tuning range is obtainable in amplifier-type devices where a smaller fraction of the energy would be stored in the tuning cavity.

When a gas atmosphere is used with spiral-beam tuning, frequency deviations of one or two orders of magnitude larger can be obtained with beam currents (driving power) one or two orders of magnitude smaller than those used in vacuum. The reasons for the improvement in tuning in a gas atmosphere are that one beam electron frees several ionization electrons as a result of ionizing collisions and that the ionization electrons interact with the electric field for a longer time due to their very low drift velocity.

Frequency changes of ± 30 per cent have been measured in gas atmospheres. These values did not necessarily represent maximum or optimum values, but rather limitations in test equipment. Fig. 19 shows gas pressure and beam current required for a given frequency deviation at both S- and X-band for the operating point $\omega_c/\omega_0 = 0.86$.

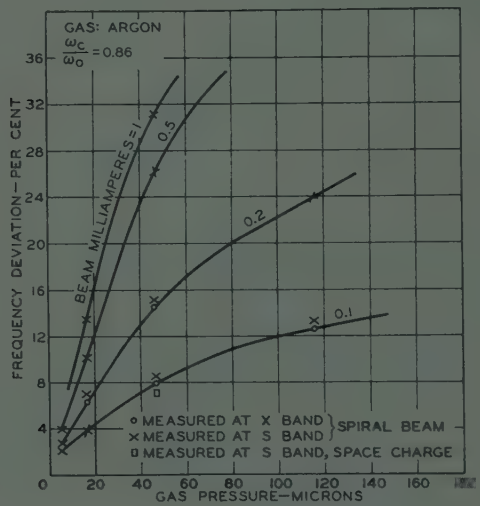


Fig. 19—Summary of results of frequency change as a function of gas pressure.

Space-charge tuning produces a tuning characteristic similar to that of spiral-beam tuning. Frequency deviations for both vacuum and gas atmosphere are in the same order of magnitude as those obtained with spiral-beam tuning. However, loading is far greater and the tuning is sensitive to filament temperature. In addition, as in all microwave tubes, the device in which the cathode is not in the interaction space has certain advantages. The spiral-beam modulation technique, therefore, is superior.

[9] H. K. Jenny, "A 7000 mc developmental magnetron for frequency modulation," *RCA Rev.*, vol. 13, pp. 202–223; June, 1952.

[10] J. S. Donal, Jr., "Modulation of continuous-wave magnetrons," *Advances in Electronics*, vol. 4, Academic Press, New York, N. Y., pp. 188–256; 1952.

List of Symbols

B_c = circuit susceptance in mhos
B_e = electronic susceptance of spiraling electron cloud in mhos
B_z = magnetic field density in webers per square meter in z-direction
C_0 = dc ridge capacitance in farads
d = separation of ridges in meters
$|e|$ = electronic charge in coulombs
E_x = rf electric field intensity in volts per meter = $E_0 e^{i\omega t}$
G_e = electronic conductance of spiraling electron cloud in mhos
$i = \sqrt{-1}$
I_0 = beam current in amperes
K = factor by which mean free path is reduced due to spiraling motion of electrons.
l = axial distance for one complete helical path in meters
L = cavity length in meters
m = electronic mass in kilograms
n = number of secondary electrons produced per primary electron
Q_L = loaded cavity Q
Q_0 = internal cavity Q
$r_{\max}$ = maximum radius of spiral beam in meters
s = length of helical path of electron in spiral-beam tuning in meters
t = time in seconds
T = temperature in degrees Kelvin
v_0 = electron entrance velocity in meters per second $(=\sqrt{2kV_0/m})$
v_x = x-component of velocity
V_ω = electronic maximum rotational energy expressed in volts
V_0 = beam voltage in volts
$\left.\begin{matrix}x\\y\\z\end{matrix}\right\}$ = Cartesian co-ordinates
θ = electron transit angle in radians
τ = electron transit time in seconds
ω = angular frequency of rf field in radians per second $(=2\pi f)$
ω_c = angular frequency of rotation of electron in radians per second $(=2\pi f_c)$
ω_0 = resonant angular frequency of untuned cavity in radians per second $(=2\pi f_0)$
ω_c/ω_0 = magnetic-field parameter
. = first derivative with respect to time
.. = second derivative with respect to time

Acknowledgment

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The Resolution of Signals in White, Gaussian Noise*

C. W. HELSTROM†

Summary—The resolution of two signals of known shapes $F_1(t)$ and $F_2(t)$ in white Gaussian noise is treated as a problem in statistical decision theory. The observer must decide which of the signals is present with a minimum probability of error. The optimum system for this decision is specified in terms of filters matched to the two signals, the outputs of which are compared. The error probability is exhibited as a function of the cross-correlation of the two signals and of the signal-to-noise ratio. If the phases of the two signals are unknown, as in radar, and if the signals are of equal strength and equal *a priori* probability, the optimum system consists of filters matched to each of the signals, each followed by a detector. The observer then bases his decision upon which of the detectors has the larger output. The probability of error is computed for this case also.

I. Introduction

THE PROBLEM of resolution can be considered from two points of view. The first is that of, for example, astronomy, in which one studies the ability of a particular instrument to produce a response which the observer can identify as the result of two sources of a certain nature rather than of one such source. Thus the resolving power of a telescope is defined in terms of the smallest angular separation of two stars, the image of which can be identified as that of two stars rather than one. Similarly, considering a conventional A-scope presentation in radar, one can ask how close two targets can be in range before their echo pips so blend as to appear to be one.

From the second point of view one studies the nature of the phenomenon rather than the instrument used to observe it. One imagines a situation in which one of two (or more) similar sources is present, and one asks an observer to identify which of them it is, permitting him to use the best system which he can design for the purpose. His observations will in general suffer interference of a statistical nature which prevents an unambiguous selection. The optimum instrument for this purpose will thus depend on the characteristics of the sources as well as on the statistical properties of the interference.

It is from the latter standpoint that we wish to study narrow-band, pulsed electrical signals such as those encountered in radar or in communications. The interference will be taken as white Gaussian noise of power N per unit of frequency. The problem will be treated by the methods of statistical decision theory by imagining that one of a class of signals is presented immersed in noise, the observer being asked to identify which member of the class it is. The observer will make this decision by picking that member of the class having the largest *a posteriori* probability calculated on the basis of the received signal $x(t)$.

In order best to understand the influence of the noise, we shall assume that one of two signals, $F_1(t)$ and $F_2(t)$, is present, the form of each being known exactly. The optimum system for deciding between the two will be derived, and the probability of error P_e per decision will be calculated. The probability of error can be used as a measure of the ambiguity of the signals; that is, it measures the extent to which the similarity of two signals causes one to be mistaken for the other when they are observed in the presence of noise.

The resolution of narrow-band, pulsed signals has been discussed by Woodward,[1] who considered the problem of determining simultaneously the range and velocity of a radar target by measuring the delay in time and the Doppler shift in frequency of a returning echo. One asks how close two such signals can be in frequency and in time of arrival before it becomes difficult to tell them apart. Woodward pointed out that their ambiguity depends on the quantity λ given by

$$\lambda = B/E$$

$$E = \tfrac{1}{2}\int_0^T |u_1(t)|^2 dt = \tfrac{1}{2}\int_0^T |u_2(t)|^2 dt$$

$$B = \tfrac{1}{2}\left|\int_0^T u_1(t)u_2^*(t)e^{-i\omega t}dt\right|, \qquad (1)$$

where $u_1(t)$, $u_2(t)$ are the complex envelopes of the signals, ω is the difference in the carrier frequencies Ω_1 and Ω_2 of the two signals (resulting e.g. from a Doppler shift), and T is the time of observation. That is, the signals are taken as

$$F_1(t) = Rl\, u_1(t)e^{i\Omega_1 t}$$
$$F_2(t) = Rl\, u_2(t)e^{i\Omega_2 t} \qquad (2)$$

and they are assumed to be of equal energies (proportional to E) and of small bandwidth compared with the carrier frequencies. The quantity λ may be called the relative cross-correlation of the two signals. Woodward[1] asserts that if the quantity λ is small, the signals can be easily distinguished, while if λ is close to unity it will be difficult to distinguish them. He discusses the form of λ for various types of signals, such as trains of pulses, frequency-modulated signals, etc.

Clearly, if there were no noise present, one could distinguish two such signals, let them differ by ever so

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[1] P. M. Woodward, "Probability and Information Theory, with Application to Radar," McGraw-Hill Book Co., Inc., New York, N. Y., p. 115; 1953.

little, by passing the input through two parallel filters, one matched to $F_1(t)$ and the other matched to $F_2(t)$. The filter giving the larger output would then determine which of the two signals had been received. (By a filter matched to a given signal we mean that filter which gives the maximum output for this signal among the class of all signals having the same energy E. Its admittance is proportional to the complex conjugate of the Fourier transform of the given signal.[2])

The effect of noise on the ambiguity of two such signals will be evaluated by the decision-theoretic approach mentioned above. In Section II it will be assumed that each of the signals $F_1(t)$ and $F_2(t)$ is known exactly. Then it turns out that the decision between them can be based on the output of a single filter, which is matched to the difference of the signals, $F_2(t)-F_1(t)$. In Section III the phases of the received signals are assumed to be completely unknown, as for instance in radar. Then one compares the outputs of detectors following parallel filters, one matched to $F_1(t)$, the other matched to $F_2(t)$. The probability of error P_e, which we define as a measure of the ambiguity of the signals, is calculated in each section under the assumption that the signals are of equal energy E and equal *a priori* probability. P_e turns out to be a function of the relative cross-correlation λ and of the signal-to-noise ratio $\rho=E/2N$. The ambiguity of signals of random phase is a minimum when $\lambda=0$, i.e., when each pair is orthogonal in the sense that the integral of their product taken over the observation interval vanishes. Thus the advantage of coding into a set of orthogonal signals in communication is indicated.

The application of decision-theoretic methods to this type of situation is not, of course, restricted to the simple cases treated here. One could imagine that the signal amplitudes are unknown, so that one is asked to distinguish between two classes of signals of the forms $A_1f_1(t)$ and $A_2f_2(t)$, in which $f_1(t)$ and $f_2(t)$ are known (except perhaps for a random phase), but in which the amplitudes are described by *a priori* probability distributions $P(A_1)$ and $P(A_2)$. In this case these *a priori* distributions would be used in computing, on the basis of the received signal $x(t)$, the *a posteriori* probability distributions of the two classes of signals, and the optimum decision procedure would be accordingly modified. In another situation, the observer may have to decide among the possibilities that either one, both, or neither of the two signals is present, the *a priori* probabilities of these alternatives being given. The optimum system would then consist of two parallel filters, each matched to one of the signals (and each followed by a linear detector if the signal phases are unknown). The output of each filter would be provided with a bias level appropriately chosen in terms of the *a priori* probabilities, and the decision would be made by comparing the two outputs with their respective bias levels. Choices among larger numbers of signals can be similarly systematized by statistical decision theory.[3]

II. Ambiguity of Signals of Known Shape

One of two signals of known waveforms $F_1(t)$ and $F_2(t)$ is received in white Gaussian noise $n(t)$, the power of which is N per unit of frequency over an input band of width W which includes and is much larger than the signal bandwidth. (If F_1, F_2, and n are taken as voltages then the quantities of power and energy are determined with respect to dissipation in a resistance of 1 ohm.) The *a priori* probabilities that F_1 and F_2 are sent are ζ and $(1-\zeta)$ respectively. The signals are observed over a period of time $0<t<T$ long enough to contain them in their entirety. Let $x(t)$ be the received signal, including the noise. Then the observer must decide between case I: $x(t)=F_1(t)+n(t)$ and case II: $x(t)=F_2(t)+n(t)$. He will pick that case for which he computes the larger *a posteriori* probability.

Let the *a posteriori* probabilities of cases I and II be p_1 and p_2 respectively. Since $T\gg 1/W$, these will be given[1]

$$p_1 = K\zeta \exp - \frac{1}{N}\int_0^T [x(t) - F_1(t)]^2 dt$$

$$p_2 = K(1-\zeta) \exp - \frac{1}{N}\int_0^T [x(t) - F_2(t)]^2 dt, \qquad (3)$$

where K is that number which makes $p_1+p_2=1$. The observer decides for case I if $p_1>p_2$, and for case II if $p_1<p_2$. This decision can as well be based on the *a posteriori* likelihood ratio, given by

$$\Lambda = p_2/p_1 = \frac{1-\zeta}{\zeta}\exp\frac{2}{N}\int_0^T [F_2(t) - F_1(t)]\, x(t)dt$$

$$\cdot \exp - \frac{1}{N}\int_0^T \{[F_2(t)]^2 - [F_1(t)]^2\} dt. \qquad (4)$$

All the factors in this expression are given except the first exponential, which is a monotonic function of its argument. Hence the decision can be based on a measurement of the quantity G given by

$$G = \int_0^T x(t)[F_2(t) - F_1(t)]dt. \qquad (5)$$

This is the cross-correlation of the received signal $x(t)$ with the difference of the two signals in question. The observer picks case I or case II accordingly as G is less than or greater than a G_0 given by

$$G_0 = \frac{N}{2}\ln\frac{\zeta}{1-\zeta} + \frac{1}{2}\int_0^T \{[F_2(t)]^2 - [F_1(t)]^2\} dt. \qquad (6)$$

The quantity G is the output at time T of a filter having an impulse response $K(\tau)$ given by

[2] J. H. Van Vleck and D. Middleton, "A theoretical comparison of the visual, aural, and meter reception of pulsed signals in the presence of noise," *Jour. Appl. Phys.*, vol. 17, p. 940; November, 1946.

[3] D. Middleton, "Modern statistical approaches to reception in communication theory," Trans. IRE, vol. PGIT-4, p. 119; September, 1954.

$$K(\tau) = F_2(T - \tau) - F_1(T - \tau), \quad 0 < \tau < T$$
$$K(\tau) = 0, \qquad \tau < 0, \qquad \tau > T, \tag{7}$$

since neither of the signals is assumed to last more than T seconds. The admittance $Y(\omega)$ of this filter is given by

$$Y(\omega) = \int_0^\infty K(\tau)e^{-i\omega\tau}d\tau = e^{-i\omega T}[\Phi_2^*(\omega) - \Phi_1^*(\omega)], \tag{8}$$

where $\Phi_1(\omega)$ and $\Phi_2(\omega)$ are Fourier transforms of $F_1(t)$ and $F_2(t)$ respectively.

Since the noise $n(t)$ is Gaussian, the quantity G is also Gaussian distributed, for it is the result of a linear operation on $x(t)$. Thus one can easily calculate the probability of error. In case I it is just the probability that $G > G_0$ when $x(t)$ is $F_1(t) + n(t)$. The average error probability is then obtained by weighting the error probabilities in the two cases in accordance with the *a priori* probabilities ζ and $(1-\zeta)$.

Let us assume that the two signals are of equal *a priori* probabilities and equal energies E, where

$$E = \int_0^T [F_1(t)]^2dt = \int_0^T [F_2(t)]^2dt. \tag{9}$$

Then $G_0 = 0$, and the error probabilities are equal in both cases. Hence the average probability of error P_e is just the probability $G > 0$ when $x(t) = F_1(t) + n(t)$. In this case mean G and variance σ^2 of G are given by[4]

$$\overline{G} = \int_0^T [F_2(t) - F_1(t)]\overline{x(t)}dt$$
$$= \int_0^T [F_2(t) - F_1(t)]F_1(t)dt = -(E - B)$$
$$\sigma^2 = \int_0^T \int_0^T [F_2(t) - F_1(t)][F_2(s) - F_1(s)]\overline{n(t)n(s)}dtds$$
$$= \frac{N}{2}\int_0^T [F_2(t) - F_1(t)]^2dt = N(E - B), \tag{10}$$

where

$$B = \int_0^T F_1(t)F_2(t)dt \tag{11}$$

is the cross-correlation of the two signals. (10) used

$$\overline{n(t)n(s)} = \frac{N}{2}\delta(t - s) \tag{12}$$

as the autocorrelation function of the noise, since this corresponds to the assumption of wideband white noise of power N per unit of frequency in the limit $W \gg 1/T$. The probability that $G > 0$ is then given by

$$P_e = (2\pi\sigma^2)^{-1/2}\int_0^\infty \exp - \frac{(G - \overline{G})^2}{2\sigma^2} dG$$
$$= \tfrac{1}{2}[1 - \Phi(\sqrt{\rho(1 - \lambda)})], \tag{13}$$

[4] The bar in $\overline{x(t)}$ refers to an ensemble average rather than to a time average, so that $\overline{x(t)} = F_1(t)$ is a function of time.

where $\Phi(x)$ is the standard error-function integral, defined by

$$\Phi(x) = \frac{2}{\sqrt{\pi}}\int_0^x e^{-t^2}dt, \quad \Phi(\infty) = 1, \tag{14}$$

and λ and ρ are given by

$$\lambda = B/E, \qquad \rho = E/2N. \tag{15}$$

In Fig. 1 we have plotted the error probability P_e as a function of λ for various values of the "signal-to-noise ratio" ρ. The ambiguity of two such signals can be defined by means of the error probability P_e. For fixed ambiguity one obtains a curve of the signal-to-noise ratio ρ versus the relative cross-correlation λ

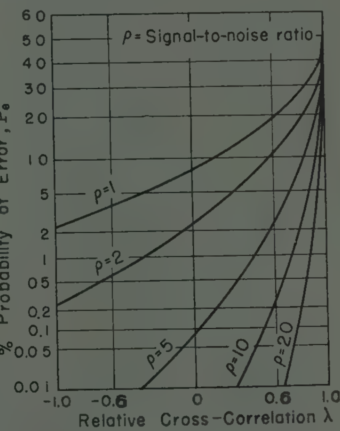

Fig. 1—Ambiguity of known signals.

which describes the effect of noise on the ambiguity of the signals. Such curves are in Fig. 2 (on following page) for error probabilities of 1, 5, and 10 per cent. One can show by means of the Schwarz inequality that $|\lambda| \leqq 1$, so that the minimum ambiguity occurs when $\lambda = -1$, i.e., when $F_2(t) = -F_1(t)$ and the two signals are 180 degrees out of phase.

III. Narrow-Band Signals of Unknown Phase

In radar systems in which the ranges of the targets are unknown *a priori* and in which no attempt is made to make successively transmitted pulses coherent, information regarding the carrier (radio-frequency) phase is lost, and one may assume it to be a uniformly distributed random variable. The same may be true in many communication systems in which coherent detection cannot be used. It is of interest to determine the ambiguity of signals in such situations.

Let us assume that the signals $F_1(t)$, $F_2(t)$ can be written as

$$F_i(t) = f_i(t)\cos[\Omega_i t + \psi_i(t) - \phi_i], \quad i = 1, 2, \tag{16}$$

where the Ω_i are the carrier frequencies and the $f_i(t)$ and $\psi_i(t)$ are the amplitude and phase modulations respectively, both the latter being of bandwidth small compared with the Ω_i. $\Omega_2 - \Omega_1 = \omega \ll \Omega_i$. The carrier phases ϕ_i are random variables distributed uniformly over their

ranges 0 to 2π. The noise is assumed to be Gaussian of power N per unit bandwidth over a range of frequencies containing both signals. Again the observer is asked to choose between two cases: (I) $x(t)=F_1(t)+n(t)$ and (II) $x(t)=F_2(t)+n(t)$; his choice is to be made in accordance with the *a posteriori* probabilities of the two cases.

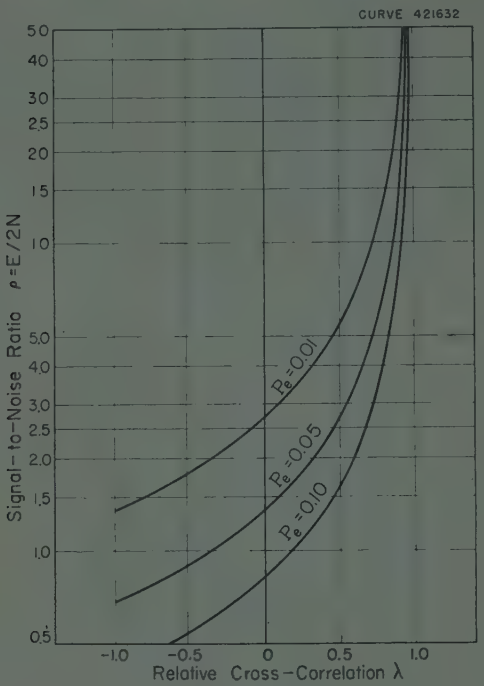

Fig. 2—Ambiguity of known signals: fixed error probability.

If the phases ϕ_1, ϕ_2 were known, the *a posteriori* probabilities would be given by (3). But since these phases are unknown, we must average the exponential factors over the distribution of the phases. This has been done by Peterson, Birdsall, and Fox[5] using a representation in terms of a sampling plan, but one can simply substitute (16) into (3), multiply by $d\phi_i/2\pi$ and integrate over $0<\phi_i<2\pi$, $i=1, 2$, using the narrow-band character of the signals to discard all but the video-frequency parts of the terms in the exponential. The *a posteriori* likelihood ratio upon which the decision is to be based then becomes

$$\Lambda = p_2/p_1 = \frac{1-\zeta}{\zeta}\exp\left(\frac{E_1-E_2}{N}\right)\frac{I_0(2R_2/N)}{I_0(2R_1/N)}, \qquad (17)$$

where ($i=1, 2$)

$$E_i = \tfrac{1}{2}\int_0^T [f_i(t)]^2 dt \qquad (18)$$

and

$$R_i^2 = X_i^2 + Y_i^2 \qquad (19)$$

with

$$X_i = \int_0^T x(t)f_i(t)\cos[\Omega_i t + \psi_i(t)]dt$$

$$Y_i = \int_0^T x(t)f_i(t)\sin[\Omega_i t + \psi_i(t)]dt. \qquad (20)$$

$I_0(x)$ is the modified Bessel function of order zero.

The observer picks case I if $\Lambda<1$ and case II if $\Lambda>1$. He could as well use the logarithm of the likelihood ratio, basing the decision on the quantity G' given by

$$G' = \ln I_0(2R_2/N) - \ln I_0(2R_1/N). \qquad (21)$$

It can be shown[5] that R_1 is the output at time T of a linear detector following a filter having the impulse response $K_1(\tau)$ given by

$$K_1(\tau) = f_1(T-\tau)\cos[\Omega_1(T-\tau)+\psi_1(T-\tau)], \quad 0<\tau<T$$
$$K_1(\tau) = 0, \quad \tau<0, \quad \tau>T. \qquad (22)$$

A filter similarly matched to $F_2(t)$ will yield, when followed by a linear detector, the quantity R_2. By using detectors having the characteristic $\ln I_0(2R/N)$, the receiver can form the quantity G' of (21). There is then a quantity G_0' depending on E_1, E_2, and ζ with which G' is compared for purposes of making a decision. Of course the detector characteristic required here is identical with that for optimum detection of pulsed signals in noise.[6]

If, however, the signals are of equal *a priori* probabilities and equal energies $E_1=E_2=E$, as we shall assume henceforth, one sees from (17) that one can simply use a linear detector (or any detector having a characteristic monotonic in R) at the output of each matched filter. One then will decide for case I if $R_1>R_2$ and for case II if $R_2>R_1$. Again the ambiguity of the two signals in noise will depend on the probability P_e of making an error in such a decision. We shall now calculate this probability as a function of the quantities λ and ρ defined in (1) and (15). The error probability P_e is the probability, given $x(t)=F_1(t)+n(t)$, that $R_2>R_1$, that is

$$P_e = \int_0^\infty dR_1\int_{R_1}^\infty p(R_1, R_2)dR_2, \qquad (23)$$

where $p(R_1, R_2)$ is the joint probability density for measuring R_1 at the output of the first and R_2 at the output of the second filter-detector combination, when the input to both is $x(t)=F_1(t)+n(t)$.

To determine the joint probability density function $p(R_1, R_2)$ one fixes the phase of $F_1(t)$ at $\phi_1=\phi$, obtaining the conditional density function $p(R_1, R_2; \phi)$. This will turn out to be independent of ϕ, so that it equals $p(R_1, R_2)$, since ϕ is completely random. The quantities

[5] W. W. Peterson, T. G. Birdsall, and W. C. Fox, "The theory of signal detectability," TRANS. IRE, vol. PGIT-4, p. 171; September, 1954. See Section 4.5.

[6] D. Middleton, "Statistical criteria for the detection of pulsed carriers in noise," *Jour. Appl. Phys.*, vol. 24, p. 371; April, 1953.

X_i, Y_i of (20) are Gaussian distributed, since they are linear combinations of Gaussian variables. The means, variances, and cross-correlations of these variables are given in the following equations. For simplicity of writing it has been assumed that there is amplitude modulation only, though the same derivation could be carried through for the general case by replacing $\Omega_i t$ by $\Omega_i t + \psi_i(t)$ everywhere.

$$\overline{X}_1 = \int_0^T [f_1(t)]^2 \cos \Omega_1 t \cos (\Omega_1 t - \phi) dt = E \cos \phi$$

$$\overline{Y}_1 = \int_0^T [f_1(t)]^2 \sin \Omega_1 t \cos (\Omega_1 t - \phi) dt = E \sin \phi$$

$$\begin{aligned}\overline{X}_2 &= \int_0^T f_1(t) f_2(t) \cos \Omega_2 t \cos (\Omega_1 t - \phi) dt \\ &= \tfrac{1}{2} \int_0^T f_1(t) f_2(t) \cos [(\Omega_2 - \Omega_1)t + \phi] dt \\ &= c_1 \cos \phi - c_2 \sin \phi = B \cos (\phi + \psi)\end{aligned}$$

$$\begin{aligned}\overline{Y}_2 &= \int_0^T f_1(t) f_2(t) \sin \Omega_2 t \cos (\Omega_1 t - \phi) dt \\ &= \tfrac{1}{2} \int_0^T f_1(t) f_2(t) \sin [(\Omega_2 - \Omega_1)t + \phi] dt \\ &= c_1 \sin \phi + c_2 \cos \phi = B \sin (\phi + \psi), \qquad (24)\end{aligned}$$

where

$$\begin{aligned}c_1 &= \tfrac{1}{2} \int_0^T f_1(t) f_2(t) \cos (\Omega_2 - \Omega_1) t \, dt = B \cos \psi \\ c_2 &= \tfrac{1}{2} \int_0^T f_1(t) f_2(t) \sin (\Omega_2 - \Omega_1) t \, dt = B \sin \psi \qquad (25) \\ B^2 &= c_1^2 + c_2^2.\end{aligned}$$

The fact that the signals are of narrow bandwidths compared with the carrier frequencies has enabled us to simplify the above integrals by keeping only the slowly varying parts of the integrands.

Because the signal energies are equal, the variances of the X_i and Y_i are all equal to σ^2, which is given by

$$\begin{aligned}\sigma^2 &= \int_0^T \int_0^T \overline{n(t)n(s)} f_1(t) f_1(s) \cos \Omega_1 t \cos \Omega_2 s \, dt ds \\ &= \frac{N}{2} \int_0^T [f_1(t)]^2 \cos^2 \Omega_1 t \, dt = NE/2, \qquad (26)\end{aligned}$$

where we have used (12). The cross-correlations are

$$\overline{(X_1 - \overline{X}_1)(Y_1 - \overline{Y}_1)} = \overline{(X_2 - \overline{X}_2)(Y_2 - \overline{Y}_2)} = 0 \qquad (27)$$

$$\begin{aligned}\overline{(X_1 - \overline{X}_1)(X_2 - \overline{X}_2)} &= \overline{(Y_1 - \overline{Y}_1)(Y_2 - \overline{Y}_2)} \\ &= \int_0^T \int_0^T \overline{n(t)n(s)} f_1(t) f_2(s) \cos \Omega_1 t \cos \Omega_2 s \, dt ds \\ &= \frac{N}{2} \int_0^T f_1(t) f_2(t) \cos \Omega_1 t \cos \Omega_2 t \, dt = N c_1/2 = k_1, \qquad (28)\end{aligned}$$

$$\begin{aligned}\overline{(X_1 - \overline{X}_1)(Y_2 - \overline{Y}_2)} &= -\overline{(X_2 - \overline{X}_2)(Y_1 - \overline{Y}_1)} \\ &= \int_0^T \int_0^T \overline{n(t)n(s)} f_1(t) f_2(s) \cos \Omega_1 t \sin \Omega_2 s \, dt ds \\ &= \frac{N}{2} \int_0^T f_1(t) f_2(t) \cos \Omega_1 t \sin \Omega_2 t \, dt = N c_2/2 = k_2. \qquad (29)\end{aligned}$$

The joint distribution of the X_1, Y_1, X_2, Y_2 is now the exponential of a quadratic form, the coefficients of the terms of which form a matrix which is the inverse of the correlation matrix of the four variables.[7] Taking the variables in the above order, the correlation matrix $\|\phi_{ij}\|$ and its inverse $\|\mu_{ij}\|$ are

$$\|\phi_{ij}\| = \begin{Vmatrix} \sigma^2 & 0 & k_1 & k_2 \\ 0 & \sigma^2 & -k_2 & k_1 \\ k_1 & -k_2 & \sigma^2 & 0 \\ k_2 & k_1 & 0 & \sigma^2 \end{Vmatrix},$$

$$\|\mu_{ij}\| = \frac{1}{A} \begin{Vmatrix} \sigma^2 & 0 & -k_1 & -k_2 \\ 0 & \sigma^2 & k_2 & -k_1 \\ -k_1 & k_2 & \sigma^2 & 0 \\ -k_2 & -k_1 & 0 & \sigma^2 \end{Vmatrix}, \qquad (30)$$

where $A = \sigma^4 - k_1^2 - k_2^2$. Thus the joint probability may be written

$$\begin{aligned}p(X_1, Y_1, X_2, Y_2) = (4\pi^2 A)^{-1} \exp &-(2A)^{-1}\{\sigma^2[(X_1 - \overline{X}_1)^2 \\ &+ (Y_1 - \overline{Y}_1)^2 + (X_2 - \overline{X}_2)^2 + (Y_2 - \overline{Y}_2)^2] \\ &- 2k_1[(X_1 - \overline{X}_1)(X_2 - \overline{X}_2) + (Y_1 - \overline{Y}_1)(Y_2 - \overline{Y}_2)] \\ &- 2k_2[(X_1 - \overline{X}_1)(Y_2 - \overline{Y}_2) - (X_2 - \overline{X}_2)(Y_1 - \overline{Y}_1)]\}. \qquad (31)\end{aligned}$$

If one now uses the above expressions for the means, making the substitutions

$$\begin{aligned}X_1 &= R_1 \cos \theta_1, & Y_1 &= R_1 \sin \theta_1, \\ X_2 &= R_2 \cos \theta_2, & Y_2 &= R_2 \sin \theta_2, \\ \theta_1' &= \theta_1 - \phi, & \theta_2' &= \theta_2 - \phi - \psi,\end{aligned}$$

$$k_1 = \mu \cos \psi, \quad k_2 = \mu \sin \psi, \quad \mu = NB/2, \qquad (32)$$

one finds the joint probability of the new variables R_1, R_2, θ_1', θ_2' to be[8]

$$\begin{aligned}p(R_1, R_2, \theta_1', \theta_2'; \phi) &= \frac{R_1 R_2}{4\pi^2 A} e^{-E/N} \exp\left\{-\frac{\sigma^2}{2A}(R_1^2 + R_2^2)\right. \\ &\left. + \frac{\mu R_1 R_2}{A} \cos (\theta_2' - \theta_1') + \frac{2R_1}{N} \cos \theta_1'\right\}, \qquad (33)\end{aligned}$$

where the factor $R_1 R_2$ comes from the Jacobian of the

[7] S. O. Rice, "The mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 23, p. 282; July, 1944; and vol. 24, p. 46; January, 1945.

[8] This derivation is formally similar to one given by D. Middleton for a different problem, "Some general results in the theory of noise through non-linear devices," *Quart. Appl. Math.*, vol. 5, p. 445; January, 1948, section 5. His eq. (5.12) can be reduced to our (33) by making the proper identification of symbols.

transformation to the new variables. Note that this expression is independent of ϕ. In order to find the joint distribution of the magnitudes R_1 and R_2, we integrate θ_1' and θ_2' over their ranges 0 to 2π. The result is

$$p(R_1, R_2) = \frac{R_1 R_2}{A} e^{-E/N} \exp\left[-\frac{\sigma^2}{2A}(R_1^2 + R_2^2)\right] \cdot I_0(\mu R_1 R_2/A) I_0(2R_1/N). \quad (34)$$

The error probability P_e is the probability that $R_2 > R_1$, i.e., (23):

$$P_e = (1-\lambda^2)e^{-2\rho} \cdot \int_0^\infty dx \int_x^\infty dy\, xy\, e^{-(x^2+y^2)/2} I_0(2x\sqrt{\rho(1-\lambda^2)}) I_0(\lambda xy), \quad (35)$$

where we have introduced the notation

$$\lambda = B/E, \; \rho = E/2N, \; x^2 = \sigma^2 R_1^2/A, \; y^2 = \sigma^2 R_2^2/A \quad (36)$$

and changed variables in the double integral. In carrying out the derivation with the inclusion of the phase modulations $\psi_i(t)$ one finds that the quantities B, E, and λ are just those given by (1), with $u_i(t) = f_i(t) \exp i\psi_i(t)$. The integral can be reduced by transformations outlined in the Appendix. The result is

$$P_e = Q(\sqrt{\rho(1-\sqrt{1-\lambda^2})}, \sqrt{\rho(1+\sqrt{1-\lambda^2})}) - \tfrac{1}{2}e^{-\rho} I_0(\rho\lambda), \quad (37)$$

where the function $Q(\alpha, \beta)$ is given by

$$Q(\alpha, \beta) = \int_\beta^\infty t\, e^{-(t^2+\alpha^2)/2} I_0(\alpha t)\, dt. \quad (38)$$

It has been tabulated by Marcum.[9]

If both λ and $\rho\lambda$ are small, it is convenient to use the series expansion

$$P_e = \tfrac{1}{2} e^{-\rho}\left[I_0(\rho\lambda) + 2\sum_{n=1}^{\infty}\left(\frac{1-\sqrt{1-\lambda^2}}{\lambda}\right)^n I_n(\rho\lambda)\right]. \quad (39)$$

For large signal-to-noise ratios, $\rho \gg 1$, and for $1-\lambda \ll 1$, (37) reduces to (13), so that in this region the loss of phase information introduces only a very small increase in the probability of error. [See Appendix, (62).] In Fig. 3 we have plotted P_e versus λ for a number of values of ρ, while in Fig. 4 are given curves of ρ versus λ for various values of P_e. By comparing these curves with those of Figs. 1 and 2 one can assess the increase in ambiguity arising when the carrier phases become uncertain. Thus it is the quantity λ, along with the signal-to-noise ratio ρ, which again determines the ambiguity of the signals, so that Woodward's contention[1] that the ambiguity of narrow-band signals depends on the relative cross-correlation λ is borne out, provided one takes account of the noise as we have done here.

[9] J. I. Marcum, "Table of Q Functions," Rand Corporation Report RM-339; January 1, 1950.

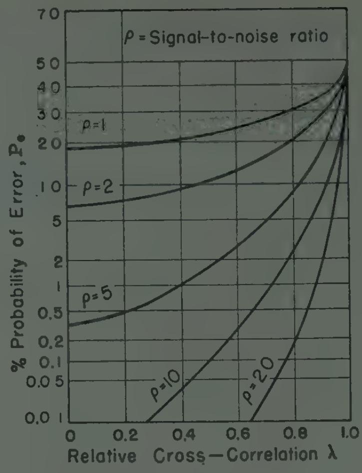

Fig. 3—Ambiguity of signals of unknown phase.

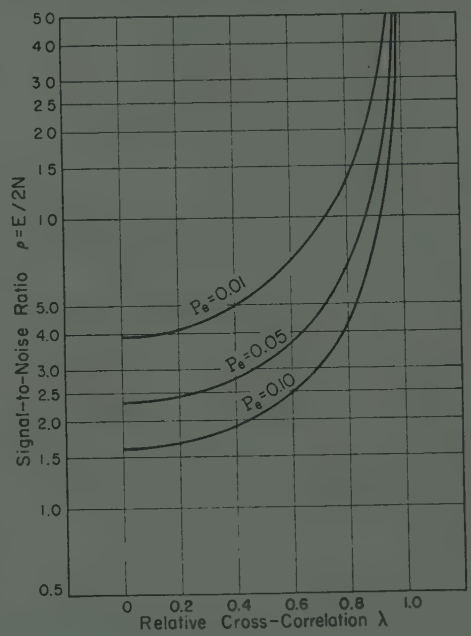

Fig. 4—Ambiguity of signals of unknown phase: fixed error probability, P_e.

IV. Conclusion

The probability of error in deciding which of two signals, $F_1(t)$ or $F_2(t)$, was sent has been computed as a function of the signal-to-noise ratio $\rho = E/2N$ and of the relative cross-correlation λ of the signals, where λ is given by (1). It has been shown that as λ approaches unity, an ever higher signal-to-noise ratio is required to keep the error probability to a pre-assigned value <0.5.

These results have a bearing upon the accuracy with which parameters of a received signal, such as its carrier frequency or its time of arrival, can be measured when noise is present.[10] Consider for example the measurement of frequency. This could be accomplished by use of a large number of filters of amplitude characteristic

[10] Cf. D. Slepian, "Estimation of signal parameters in the presence of noise," Trans. IRE, vol. PGIT-3, p. 68; March, 1954.

matching the pulse envelope, the pass-frequencies Ω_i spaced more or less uniformly over the band of expected signal carrier frequencies. That filter yielding the maximum output would determine the signal frequency to an accuracy given by the frequency spacing between adjacent filters. Now there would be little point to placing the filters so close together that the noise would introduce a large probability of error in the decision as to which filter output was the largest. If one considers the filters pairwise, the results of this paper enable one to determine the probability that an adjacent filter, of pass frequency Ω_{i+1}, say, will have a larger output than that of pass frequency Ω_i, when the frequency of the signal was really Ω_i. By setting a limit to this probability P_e one can determine the overlap λ as a function of the expected signal-to-noise ratio ρ, using the curves of Figs. 2 or 4.

Suppose for example that one expects pulses of Gaussian envelope $u(t)$ given by

$$u(t) = A \exp\left[-\tfrac{1}{2}\alpha^2(t - \tfrac{1}{2}T)^2\right], \tag{40}$$

where α is roughly the bandwidth of the pulse. If we assume that the observation time is long compared with α^{-1}, the relative cross-correlation λ is given by (1) to be

$$\lambda = e^{-\omega^2/4\alpha^2} \tag{41}$$

when the time of arrival of the pulses is the same, but when the frequency separation (corresponding to the difference of the pass frequencies of adjacent matched filters) is $\Omega_{i+1} - \Omega_i = \omega$. If we use for simplicity the results of Section II, which assume the signals completely known, we find that the dependence of signal-to-noise ratio upon λ for fixed error probability P_e is given by

$$\rho(1 - \lambda) = k^2, \tag{42}$$

where k is a constant such that

$$P_e = \tfrac{1}{2}[1 - \Phi(k)]. \tag{43}$$

Now for λ near unity, (41) is, approximately,

$$\omega = 2\alpha\sqrt{-\ln\lambda} \cong 2\alpha\sqrt{1-\lambda} \tag{44}$$

so that

$$\omega \cong 2k\alpha/\sqrt{\rho}. \tag{45}$$

This implies that the minimum resolvable frequency difference, i.e., the minimum reasonable difference between the pass frequencies of adjacent filters, is proportional to the signal bandwidth and inversely proportional to the square root of the signal-to-noise ratio. This essentially is the limitation upon the accuracy with which the frequency of such a signal can be determined. Of course, to make better use of such a system one should use the outputs of all the filters of the array, computing from them the *a posteriori* probability distribution of the input signal frequency. The width of this distribution would indicate the expected error in a frequency determination by this means.

Appendix

In order to evaluate the integral of (35), we start with the definition (38) of the function $Q(\alpha, \beta)$. By integration by parts[11] one can show, for $\alpha < \beta$,

$$Q(\alpha, \beta) = e^{-(\alpha^2+\beta^2)/2} \sum_{n=0}^{\infty} (\alpha/\beta)^n I_n(\alpha\beta). \tag{46}$$

Now we use the integral representation of the modified Bessel functions $I_n(x)$:

$$I_n(x) = \frac{1}{2\pi}\int_0^{2\pi} \cos n\theta \, e^{x\cos\theta} d\theta. \tag{47}$$

Substituting into (46) and interchanging the order of summation and integration we get

$$\begin{aligned} Q(\alpha,\beta) &= \frac{1}{2\pi} e^{-(\alpha^2+\beta^2)/2} \int_0^{2\pi} e^{\alpha\beta\cos\theta} \sum_{n=0}^{\infty} (\alpha/\beta)^n \cos n\theta \, d\theta \\ &= \frac{1}{2\pi} e^{-(\alpha^2+\beta^2)/2} \\ &\quad \cdot \int_0^{2\pi} \frac{1 - (\alpha/\beta)\cos\theta}{1 - 2(\alpha/\beta)\cos\theta + (\alpha/\beta)^2} e^{\alpha\beta\cos\theta} d\theta \\ &\qquad (\alpha < \beta). \end{aligned} \tag{48}$$

Now (35) can be written using (38) as

$$P_e = (1 - \lambda^2)e^{-2\rho} \cdot \int_0^{\infty} x e^{-x^2(1-\lambda^2)/2} I_0(2x\sqrt{\rho(1-\lambda^2)}) Q(\lambda x, x) dx. \tag{49}$$

Substituting from (48) for $Q(\lambda x, x)$, we get

$$P_e = \frac{(1-\lambda^2)e^{-2\rho}}{2\pi} \int_0^{2\pi} \frac{1 - \lambda\cos\theta}{1 - 2\lambda\cos\theta + \lambda^2} d\theta \cdot \int_0^{\infty} x e^{-x^2(1-\lambda\cos\theta)} I_0(2x\sqrt{\rho(1-\lambda^2)}) dx. \tag{50}$$

Now we use the formula

$$\int_0^{\infty} x e^{-a^2x^2} I_0(bx) dx = \frac{e^{b^2/4a^2}}{2a^2} \tag{51}$$

(from which one can show that $Q(\alpha, 0) = 1$). (50) then becomes, with $a^2 = 1 - \lambda\cos\theta$, $b = 2\sqrt{\rho(1-\lambda^2)}$,

$$P_e = \frac{(1-\lambda^2)e^{-2\rho}}{4\pi} \int_0^{2\pi} \frac{d\theta}{1 - 2\lambda\cos\theta + \lambda^2} \cdot \exp\left[\frac{\rho(1-\lambda^2)}{1 - \lambda\cos\theta}\right]. \tag{52}$$

We now make the change of variable given by

[11] J. I. Marcum, *A Statistical Theory of Target Detection by Pulsed Radar*, Math. App., Rand Corp. Report RM-753; 1948.

$$\cos \theta = \frac{\lambda + \cos \phi}{1 + \lambda \cos \phi}, \tag{53}$$

where the range $0<\theta<2\pi$ corresponds to $0<\phi<2\pi$, obtaining after some labor

$$P_e = \frac{\sqrt{1-\lambda^2}\, e^{-\rho}}{4\pi} \int_0^{2\pi} \frac{e^{\rho\lambda \cos \phi} d\phi}{1 - \lambda \cos \phi}. \tag{54}$$

Now in (48) let us put $\alpha/\beta = \mu$, whereupon we can break up the integral as follows:

$$Q(\alpha, \beta) = \frac{1}{4\pi} e^{-(\alpha^2+\beta^2)/2} \int_0^{2\pi} e^{\alpha\beta \cos \theta} d\theta \left[1 + \frac{1-\mu^2}{1-2\mu\cos\theta+\mu^2}\right]$$

$$= \tfrac{1}{2} e^{-(\alpha^2+\beta^2)/2} \left[I_0(\alpha\beta) + \frac{1-\mu^2}{2\pi} \int_0^{2\pi} \frac{e^{\alpha\beta\cos\theta} d\theta}{1-2\mu\cos\theta+\mu^2}\right] \tag{55}$$

so that

$$\frac{1}{2\pi}\int_0^{2\pi} \frac{e^{\alpha\beta\cos\theta} d\theta}{1-2\mu\cos\theta+\mu^2} = (1-\mu^2)^{-1}\left[2Q(\alpha,\beta)e^{(\alpha^2+\beta^2)/2} - I_0(\alpha\beta)\right]. \tag{56}$$

Thus we can evaluate (54) by putting

$$\lambda = \frac{2\mu}{1+\mu^2}, \quad \mu = \frac{1-\sqrt{1-\lambda^2}}{\lambda}, \quad \sqrt{1-\lambda^2} = \frac{1-\mu^2}{1+\mu^2}$$

$$\alpha = \sqrt{\rho\lambda\mu} = \sqrt{\rho(1-\sqrt{1-\lambda^2})},$$

$$\beta = \sqrt{\rho\lambda/\mu} = \sqrt{\rho(1+\sqrt{1-\lambda^2})}, \tag{57}$$

whereupon (54) becomes

$$P_e = \frac{(1-\mu^2)e^{-\rho}}{4\pi}\int_0^{2\pi} \frac{(e^{\rho\lambda})^{\cos\theta} d\theta}{1-2\mu\cos\theta+\mu^2}$$

$$= \tfrac{1}{2} e^{-\rho}\left[2Q(\alpha,\beta)e^{(\alpha^2+\beta^2)/2} - I_0(\alpha\beta)\right]$$

$$= Q(\alpha,\beta) - \tfrac{1}{2} e^{-\rho} I_0(\rho\lambda) \tag{58}$$

which was to be proved. The series expansion of (39) comes directly from (46).

When α and β are large, one can obtain an asymptotic evaluation of $Q(\alpha, \beta)$ by using the asymptotic form of the function $I_0(x)$:[10]

$$I_0(x) \cong \frac{e^x}{\sqrt{2\pi x}}. \tag{59}$$

Putting this into the integral (38), and noticing that most of the contribution to the integral comes from the region in which $y \sim \alpha$, one obtains

$$Q(\alpha,\beta) \cong \int_\beta^\infty \frac{y}{\sqrt{2\pi\alpha y}} e^{-(y-\alpha)^2/2} dy$$

$$\sim \frac{1}{2\pi}\int_\beta^\infty e^{-(y-\alpha)^2/2} dy$$

$$= \tfrac{1}{2}\left[1 - \Phi\left(\frac{\beta-\alpha}{\sqrt{2}}\right)\right]. \tag{60}$$

Now for $1-\lambda \ll 1$, one can write α and β approximately as

$$\alpha \doteq \sqrt{\rho}(1-\sqrt{(1-\lambda)/2}), \quad \beta \doteq \sqrt{\rho}(1+\sqrt{(1-\lambda)/2}). \tag{61}$$

In addition, the second term of (58) becomes negligible, so that one finally obtains the approximate result

$$P_e \cong \tfrac{1}{2}\left[1 - \Phi(\sqrt{\rho(1-\lambda)})\right]. \tag{62}$$

Comparison of the curves of Figs. 1 and 3 shows that this is a good approximation when $\rho \gtrsim 10$, $\lambda \gtrsim 0.5$.

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Automatic Gain Control of Transistor Amplifiers*

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Summary—**Since transistor small signal parameters are functions of the dc emitter current (I_e) and of the dc collector voltage (V_c), gain control can be achieved by varying either I_e or V_c. The gain decreases with decreasing I_e or V_c.**

Using the series-parallel representation, the parameter most sensitive to I_e-variations is h_{11}, whereas V_c-variations affect h_{12} and h_{22} considerably. In common emitter configuration changes of h_{21} are also important. A study of the dependance of the h_{ij} on the dc operating point explains the nature of gain variations with I_e and V_c.

Satisfactory AGC circuits have been built using either I_e- or V_c-control. The control power required is very small if I_e or V_c are controlled indirectly by varying the base current. Since I_e or V_c are decreased considerably in the presence of strong input signals, the problem of distortion must be given serious consideration. Due to the variation of transistor driving point impedances, AGC may result in changes of the bandpass characteristic of tuned amplifiers.

The gain of transistor converters and oscillator-converters can be controlled by conventional or special techniques.

I. Introduction

THE POSSIBILITY of controlling the gain of amplifiers is very important in many electronic systems. Methods achieving gain control in vacuum tube amplifier circuits are well-known. These methods are based on the fact that the transconductance, and consequently, the amplification of a vacuum tube are functions of its grid bias.

Automatic gain control circuits using transistors have been described by Blecher,[1] Barton,[2] Stern and Raper.[3,4] Blecher discusses circuits using the common base configuration. The other papers describe broadcast receivers employing AGC.

The purpose of this paper is to review the theoretical aspects of transistor gain control and some of the principles useful in the design of transistor AGC circuits. The investigations leading to this paper were mainly concerned with the gain control of amplifiers designed to operate in the 100 kc to 2 mc frequency range, but it is believed that the conclusions can be considered valid for amplifiers designed for different frequencies.

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[1] F. H. Blecher, "Automatic gain control of junction transistor amplifiers," *Proc. NEC*, vol. 9, pp. 731–737; 1953.

[2] L. E. Barton, "An experimental transistor personal broadcast receiver," PROC. IRE, vol. 42, pp. 1062–1066; July, 1954.

[3] A. P. Stern and J. A. A. Raper, "Transistor AM broadcast receivers," *1954 IRE Convention Record*, part 7, "Broadcasting and Television," pp. 8–14.

[4] A. P. Stern and J. A. A. Raper, "Transistor broadcast receivers," *Elec. Eng.*, vol. 73, pp. 1107–1112; December, 1954.

II. Transistor Behavior and DC Operating Point

Small Signal Parameters and Cain

Using the series-parallel representation, the behavior of the transistor is described by:

$$\begin{aligned} E_1 &= h_{11}I_1 + h_{12}E_2 \\ I_1 &= h_{21}I_1 + h_{22}E_2. \end{aligned} \tag{1}$$

The small signal parameters h_{ij} vary with the frequency[5,6] and are, of course, different for the three transistor configurations (common base, emitter and collector). If the transistor (Fig. 1) is terminated by a

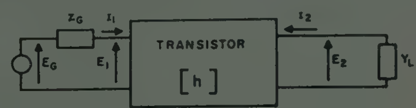


Fig. 1—Schematic representation of a transistor amplifier.

source impedance $Z_G(Z_G = R_G + jX_G)$ and a load admittance $Y_L(Y_L = G_L + jB_L)$, the transducer gain G of the transistor can be expressed as:

$$G = \frac{4R_G G_L \,|\, h_{21} \,|^2}{|\,(h_{11} + Z_G)(h_{22} + Y_L) - h_{12}h_{21}\,|^2}. \tag{2}$$

The gain G of (2) is defined as the ratio of the power delivered to the load to the available power of the source connected to the input terminals of the transistor.

Gain control of the transistor is possible because, as will be shown in the following discussion, the parameters h_{ij} depend on the dc operating point, i.e., on the dc emitter current (I_e) and the dc collector-voltage (V_c). Therefore, there are two basic methods of transistor gain control:

1. Emitter current or I_e-control
2. Collector voltage or V_c-control.

The two methods apply in different regions of the collector plane: I_e-control applies at "normal" values of V_c (several volts) and small values of I_e, whereas V_c-control involves "normal" values of I_e (order of milliampere) and small values of V_c.

The h_{ij} being frequency dependent and complex and the functions $h_{ij} = F(I_e, V_c)$ rather involved, an exact analytic treatment of the gain as a function of the dc operating point using (2) is hardly practical, especially

[5] J. M. Early, "Design theory of junction transistors," *Bell Sys. Tech. Jour.*, vol. 32, pp. 1271–1312; November, 1953.

[6] R. L. Pritchard, "Frequency variation of junction transistor parameters," PROC. IRE, vol. 42, pp. 786–799; May, 1954.

at higher frequencies. The effect of I_e and V_c on gain and other performance characteristics of transistor amplifiers can, however, be described qualitatively by analyzing the theoretical properties and observing the experimental behavior of the h_{ij} as functions of I_e and V_c.

In the following discussion, $h_{ij,b}$ designate the h-parameters of the *common base* transistor configuration whereas $h_{ij,e}$ refer to the *common emitter* circuit. Wherever it is necessary to distinguish between collector-to-base and collector-to-emitter voltages, the former is designated by V_{cb} and the latter by V_{ce}.

Common Base Parameters: Theory

The common base small signal admittance parameters of the transistor have been calculated by Early[5] in terms of the physical properties of the device. Early has solved the one-dimensional diffusion equation approximately valid for an "ideal" transistor and has added to the ideal model several circuit elements representing the deviation of a "real" transistor from the ideal one.

In the case of gain controlled amplifiers I_e may be reduced to a few microamperes and V_{cb} to a few millivolts (V_{cb} may change sign). Consequently, I_e may be of the order of magnitude of the emitter or collector reverse currents and V_{cb} is not necessarily larger than kT/q. Early's solution of the diffusion equation can be written to include terms which may be of importance at small values of I_e and V_{cb}. The resulting admittance parameters are transformed into series-parallel parameters h_{ij}', yielding an equivalent circuit of the "ideal" transistor. The ideal model is assumed to have unity emitter efficiency and collector multiplication and can be completed by adding the "base spreading impedance" z_b' and collector barrier capacitance C_b (Fig. 2). (Pritchard and Coffey[7] have shown that the base spreading "resistance" is complex for rate grown n-p-n transistors.)

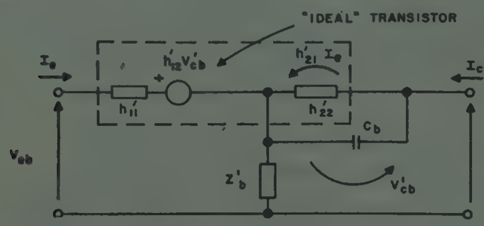


Fig. 2—"Real" transistor.

The collector-to-base leakage resistance is negligible at medium and high frequencies under consideration.

At higher frequencies h_{12}' is small as compared to the feedback due to z_b' and the parameters $h_{ij,b}$ of the common base transistor are related to those of the "ideal" transistor h_{ij}' by the following approximate relations:

[7] R. L. Pritchard and W. N. Coffey, "Small signal parameters of grown-junction transistors at high frequencies," *1954 IRE Convention Record*, part 3, "Electron Devices and Component Parts," pp. 89–98.

$$h_{11,b} = h_{11}' + z_b'(1 + h_{21}') \tag{3}$$

$$h_{12,b} = (h_{22}' + j\omega C_b)z_b' \tag{4}$$

$$h_{21,b} = -\alpha = h_{21}' \tag{5}$$

$$h_{22,b} = h_{22}' + j\omega C_b. \tag{6}$$

Writing out the h_{ij}' explicitly, the $h_{ij,b}$ can be written as functions of I_e and V_{cb}:

$$h_{11,b} = \frac{\tanh sw/L}{as \tanh w/L} \frac{1}{I_e + I_{e0}} + z_b'\left(1 - \frac{1}{\cosh sw/L}\right) \tag{7}$$

$$h_{12,b} = \left\{\left[aI' + \frac{1}{L}\frac{\partial w}{\partial V_{cb}}\left(\frac{1}{\cosh w/L} I_e + I_{c0}\right)\right] s \cdot \tanh sw/L + j\omega C_b\right\} z_b' \tag{8}$$

$$h_{21,b} = -\frac{1}{\cosh sw/L} \tag{9}$$

$$h_{22,b} = \left[aI' + \frac{1}{L}\frac{\partial w}{\partial V_{cb}}\left(\frac{1}{\cosh w/L} I_e + I_{c0}\right)\right] s \cdot \tanh sw/L + j\omega C_b. \tag{10}$$

I_{e0} is the emitter reverse current, the collector being biased at V_{cb}; I_{c0} is the collector reverse current with open emitter and I' is a quantity having the dimension of a current. For p-n-p transistors:

$$I_{e0} = \frac{qD_p p_n}{L \sinh w/L}(e^{aV_{cb}} - 1 + \cosh w/L) \tag{11}$$

$$I_{c0} = \frac{qD_p p_n}{L}(1 - e^{aV_{cb}}) \tanh w/L \tag{12}$$

$$I' = \frac{qD_p p_n}{L} e^{aV_{cb}}. \tag{13}$$

The symbols used in (7) to (13) are those of Early and have the following significance:

- $a = q/kT$ (approximately 40 at room temperature);
- D_p = diffusion constant for holes;
- p_n = equilibrium concentration of holes in n-type base region;
- L = diffusion length of holes;
- w = base-layer thickness;
- $s = (1+j\omega\tau)^{1/2}$, τ being the lifetime of holes in the base region.

Eq. (7) shows that $h_{11,b}$ has a component which is inversely proportional to (I_e+I_{e0}) and a component proportional to $(1+h_{21,b})$. Due to the decrease of $\alpha = -h_{21,b}$ with decreasing I_e at low values of I_e, $(1+h_{21,b})$ increases under the same conditions. Consequently, with decreasing I_e, $h_{11,b}$ increases. $h_{11,b}$ is, however, hardly affected by changes in V_{cb}.

According to (8) and (10), the dependence of $h_{12,b}$ and $h_{22,b}$ on the dc parameters is rather complicated.

Both $h_{12,b}$ and $h_{22,b}$ have a component varying linearly with I_e and all components are sensitive to variations V_{cb}, due to the fact that both $\partial w/\partial V_{cb}$ and C_b increase with decreasing V_{cb}. The major portion of $h_{12,b}$ and $h_{22,b}$ in practical transistors is due to C_b and, C_b being independent of I_e; the variation of these parameters with I_e is not as strong as their variation with V_{cb}.

In (9) $h_{21,b}$ is independent of I_e and V_{cb}, but it is well known that $h_{21,b}=-\alpha$ does decrease at low values of I_e[8] and in the neighborhood of zero V_{cb}.

In the case of transistors with small base layer thickness the dependence of $h_{ij,b}$ on V_{cb} is particularly complicated because of the nonnegligible variation of the base layer thickness with V_{cb}.[9]

The variation of the h-parameters with I_e can be summarized schematically by the following qualitative relationships:

$$h_{11,b} \cong A_1 \frac{1}{I_e + I_{e0}} + z_b'(1-\alpha) \qquad (14)$$

$$h_{12,b} \cong [A_2(\alpha_0 I_e + I_{c0}) + A_3]z_b' \qquad (15)$$

$$h_{21,b} = -\alpha \qquad (16)$$

$$h_{22,b} \cong A_2(\alpha_0 I_e + I_{c0}) + A_3. \qquad (17)$$

The A_i are complex constants and α_0 is the low frequency value of α. A_3 is usually the prevailing term in (15) and (17) and consequently, the parameter most sensitive to I_e variations is $h_{11,b}$. Therefore, one can state that, in terms of small signal parameter variations, *emitter current type gain control is due principally to variations of* $h_{11,b}$.

In the case of varying V_{cb}, both $\partial w/\partial V_{cb}$ and C_b are functions of V_{cb}.[5,10] In general:

$$\frac{\partial w}{\partial V_{cb}} = k_1 V_{cb}^{-m} \qquad (18)$$

$$C_b = k_2 V_{cb}^{-n}. \qquad (19)$$

The magnitude of the exponents m and n depends on the nature of the collector junction. For graded junctions $m=\frac{2}{3}$ and $n=\frac{1}{3}$, whereas for step junctions $m=n=\frac{1}{2}$. Both cases are idealized: in practice the exponents will be close to $\frac{1}{2}$.

The h-parameters for varying V_{cb} can be written in first approximation as:

$$h_{11,b} \cong B_1 + z_b'(1-\alpha) \qquad (20)$$

$$h_{12,b} \cong [B_2 e^{aV_{cb}} + B_3 V_{cb}^{-m} + B_4 V_{cb}^{-n}]z_b' \qquad (21)$$

$$h_{21,b} = -\alpha \qquad (22)$$

$$h_{22,b} \cong B_2 e^{aV_{cb}} + B_3 V_{cb}^{-m} + B_4 V_{cb}^{-n}. \qquad (23)$$

[8] W. M. Webster, "On the variation of junction-transistor current-amplification factor with emitter current," Proc. IRE, vol. 42, pp. 914–920; June, 1954.

[9] D. Haneman, "Expression for the "α-cut-off frequency in junction-transistors," Proc. IRE, vol. 42, pp. 1808–1809; December, 1954.

[10] J. M. Early, "Effects of space-charge layer widening in junction-transistors," Proc. IRE, vol. 40, pp. 1401–1406; November, 1952.

These expressions show that $h_{12,b}$ and $h_{22,b}$ are the parameters most sensitive to variations of V_{cb}. Therefore, in terms of small signal parameter variation, *collector voltage type gain control is due principally to variations of* $h_{12,b}$ *and* $h_{22,b}$.

Common Base Parameters: Experimental Results

The common base h-parameters of a typical General Electric rate grown n-p-n transistor (Type 2N78) were measured as functions of I_e and V_{cb} at one megacycle.

$h_{11,b}$ is considered as the parallel connection of a resistance $r_{11,b}$ and a (negative) capacitance $C_{11,b}$ (Fig. 3). Both resistive and reactive components of $h_{11,b}$ increase if I_e is decreasing. The phase angle of $h_{11,b}$ decreases as I_e is decreasing and reverses itself ($C_{11,b}$ becomes positive) at a small value of I_e (this part of the curve is not shown in Fig. 3). The phase reversal is due to the fact that at low values of I_e, the first component of $h_{11,b}$ in (7) is prevailing and this component is capacitive. $h_{11,b}$ depends only to a very moderate extent on V_{cb}.

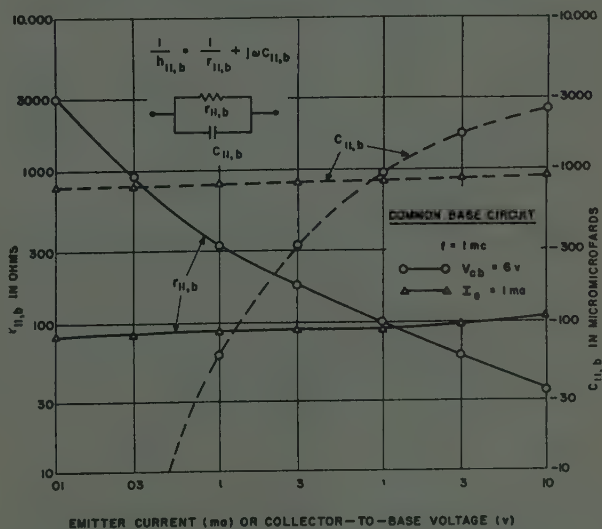


Fig. 3—Components of $h_{11,b}$ as functions of emitter current and collector-to-base voltage.

$h_{12,b}$ (Fig. 4, next page) does not vary strongly with I_e, small measured variation due mainly to I_e-dependence of z_b'. The variation of $h_{12,b}$ with V_{cb} is considerable, as can be expected. $h_{12,b}$ increases if V_{cb} is decreasing but its phase remains unaffected, z_b' being independent of V_{cb}.[11]

Variation of $h_{22,b}$ (Fig. 5, next page) with I_e and V_{cb} is analogous to $h_{12,b}$ as can be expected from the similarity of (8) and (10). Effect of I_e is small, whereas that of V_{cb} is considerable. Both components of $h_{22,b}$ ($g_{22,b}$ and $C_{22,b}$) increase if V_{cb} is decreasing. The slope $C_{22,b}$ versus V_{cb} is comparable to that of $h_{12,b}$ versus V_{cb} and is close to $(-1/2)$.

[11] Measurements of the phases of $h_{12,b}$ and $h_{12,e}$ involved considerable errors and the corresponding curves may not be representative.

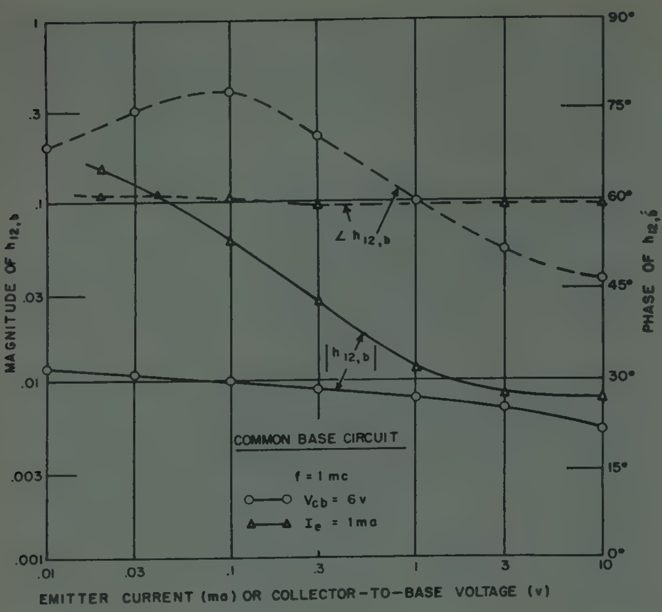


Fig. 4—$h_{12,b}$ as a function of emitter current and collector-to-base voltage.

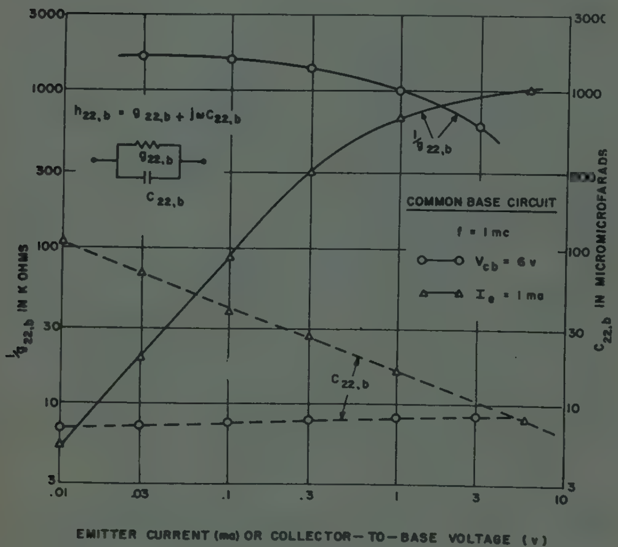


Fig. 5—Components of $h_{22,b}$ as functions of emitter current and collector-to-base voltage.

The magnitude of $h_{21,b} = -\alpha$ decreases with decreasing I_e (Fig. 6) in the region of small values of $I_e (I_e < 0.3$ ma). $h_{21,b}$ also decreases slightly with decreasing V_{cb}.

The reduction of $h_{21,b}$ at 1 mc with decreasing I_e or V_{cb} is due both to the decrease of the low frequency value of $h_{21,b}$ as well as to the decrease of the $h_{21,b}-$(or $\alpha-$) cutoff frequency. The latter manifests itself in the notable increase of the phase angle of $h_{21,b}$.

Common Emitter Parameters

The approximate relationships between common emitter parameters $h_{ij,e}$ and common base parameters $h_{ij,b}$ are:

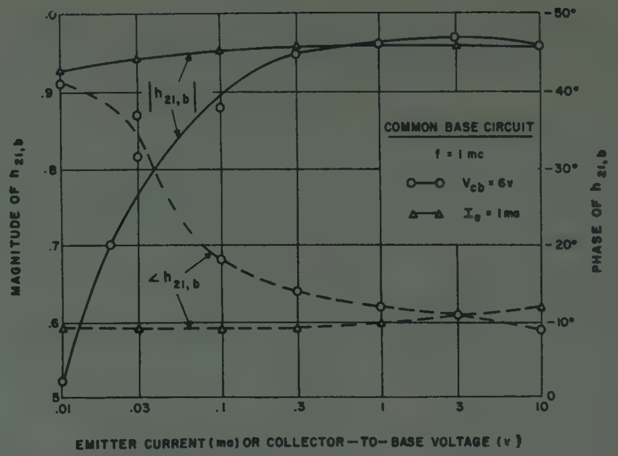


Fig. 6—$h_{21,b}$ as function of emitter current and collector-to-base voltage.

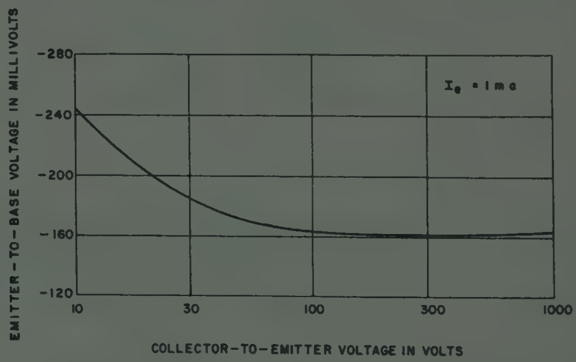


Fig. 7—Emitter-to-base versus collector-to-emitter voltage at constant emitter current.

$$h_{11,b} \cong h_{11,b}/(1 + h_{21,b}) \qquad (24)$$

$$h_{12,e} \cong h_{11,b}h_{22,b}/(1 + h_{21,b}) - h_{12,b} \qquad (25)$$

$$h_{21,b} \cong - h_{21,b}/(1 + h_{21,b}) \qquad (26)$$

$$h_{22,e} \cong h_{22,b}/(1 + h_{21,b}). \qquad (27)$$

With exception of $h_{12,e}$

$$|h_{ij,e}| \cong |h_{ij,b}/(1 - \alpha)|, \qquad (28)$$

where $1/(1-\alpha)$ decreases strongly with decreasing I_e or V_c.

In Figs. 8 to 11, the $h_{ij,e}$ are plotted as functions of I_e and V_{ce}. V_{ce} has been chosen as independent variable rather than V_{cb} for the common emitter case. By subtracting the emitter-to-base voltage from V_{ce} one obtains V_{cb} (Fig. 7).

Both resistive and reactive components of $h_{11,e}$ increase as I_e is decreasing (Fig. 8). The effect of V_{ce} variation is negligible until V_{ce} reaches values of approximately 100 mv. In this region, the collector diode is forward biased and $h_{11,e}$ decreases as V_{ce} is decreased.

$h_{12,e}$ increases if I_e is decreasing (Fig. 9) due to the increase of $h_{11,b}$ in (25). $h_{12,e}$ also increases with V_{ce}, the variation being largest in the region of forward biased collector junction.

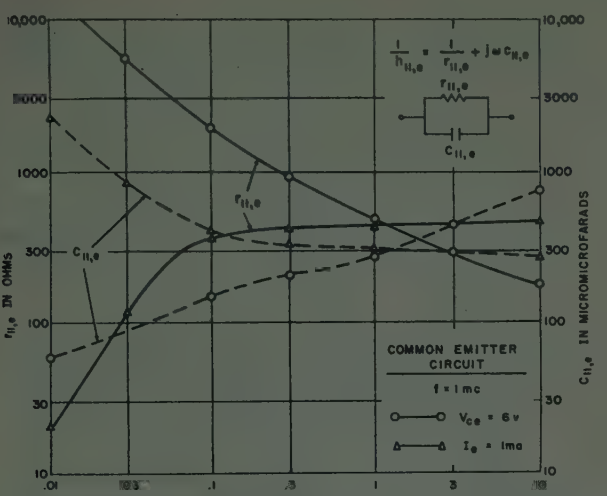


Fig. 8—Components of $h_{11,e}$ as functions of emitter current and collector-to-emitter voltage.

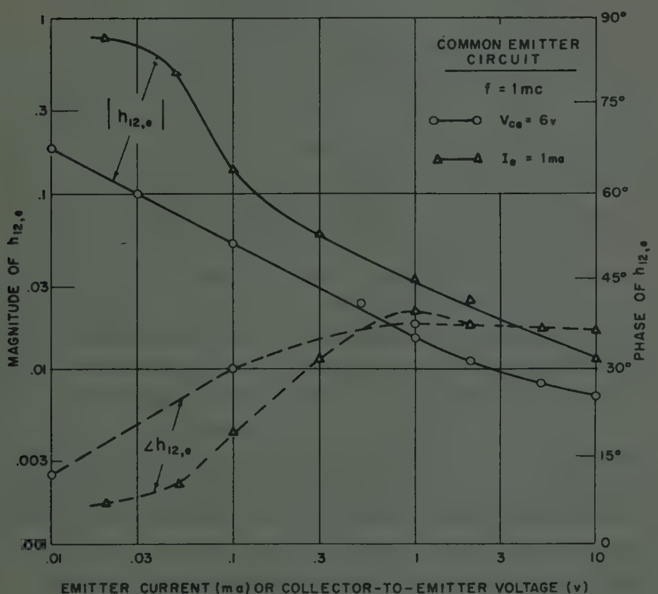


Fig. 9—$h_{12,e}$ as a function of emitter current and collector-to-emitter voltage.

$h_{21,e} \cong \alpha/(1-\alpha)$ is, of course, more sensitive to variations of I_e and V_c than $h_{21,b} = -\alpha$ (Fig. 10). The variation of $h_{22,e}$ is qualitatively similar to $h_{22,b}$ (Fig. 11).

The measured curves show that in common emitter configuration, just as in the case of the common base stage, emitter current type gain control is achieved by varying $h_{11,e}$, whereas if collector voltage type gain control is used, $h_{12,e}$ and $h_{22,e}$ are mainly responsible for gain variation. In common emitter configuration, however, the variation of $h_{21,e}$, which is in the numerator of (2), helps the gain control process considerably.

The Gain as Function of I_e and V_c.

Considering (2) for the gain and the dependence of the

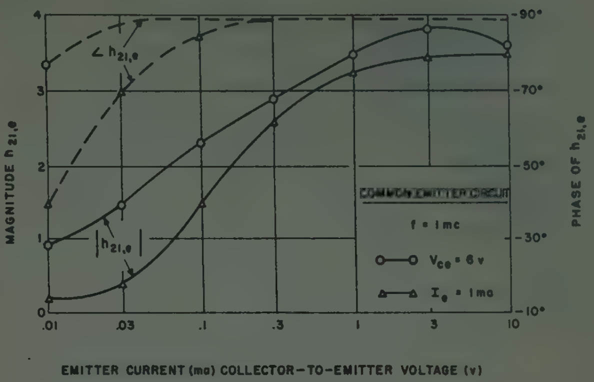


Fig. 10—$h_{21,e}$ as function of emitter current and collector-to-emitter voltage.

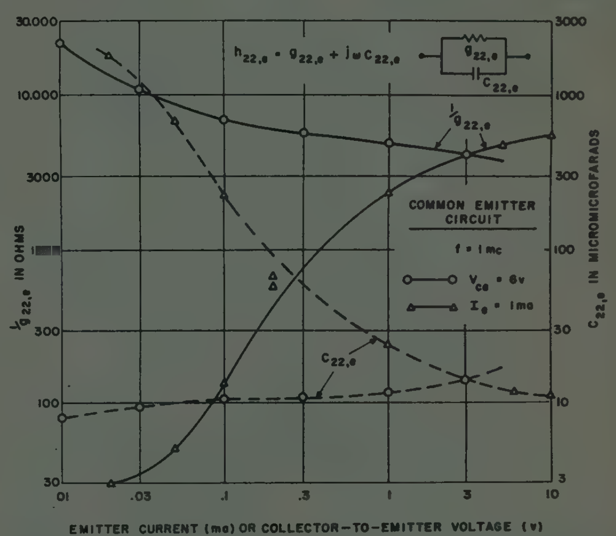


Fig. 11—Components of $h_{22,e}$ as functions of emitter current and collector-to-emitter voltage.

h-parameters on I_e, it is easy to see that in the case of I_e-control, at small values of I_e, the gain is approximately proportional to $1/|h_{11}|^2$. Due to the nature of the variation of h_{11} with I_e this means that the gain will be reduced by almost 20 db per decade decrease of I_e.

In the case of V_c-control the variation of h_{12} and h_{22} is most important. The exponent $(-\frac{1}{2})$ leads to a gain variation of approximately 10 db per decade variation of V_c at low values of V_c.

The actual dependence of the transducer gain of a General Electric rate grown n-p-n transistor on I_e and V_c has been measured at 500 kc (Figs. 12 and 13). The terminations were resistive: $R_G = 500\Omega$ and $R_L = 5{,}000\ \Omega$. The curves show that considerable control action occurs in the region of small values of I_e and V_c. (By reducing I_e and V_c beyond the values shown on the diagram, gain reductions exceeding 35 db per stage can be achieved.) The agreement with the theoretical gain variation is satisfactory: gain decrease of 15 to 18 db per decade decrease in I_e is measured in the case of I_e-control,

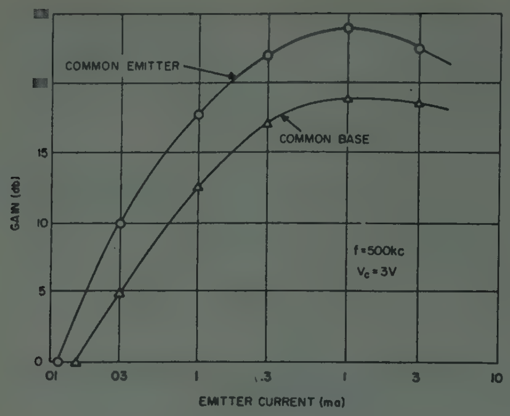


Fig. 12—Transistor gain as a function of the emitter current.

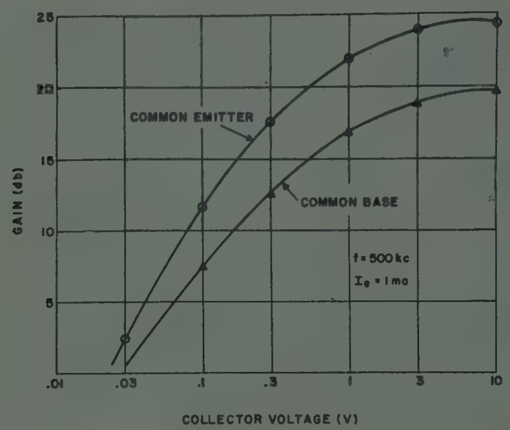


Fig. 13—Transistor gain as a function of the collector voltage.

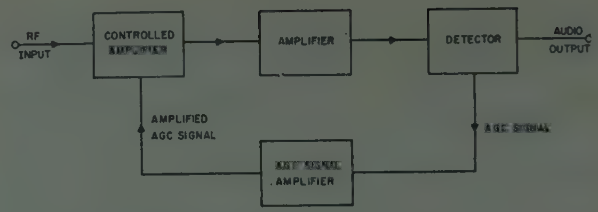
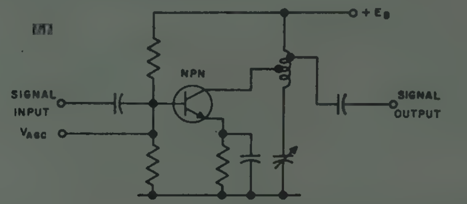


Fig. 14—AGC system with additional amplifier in feedback loop.



I. Control of tuned amplifier with AGC voltage applied to base.

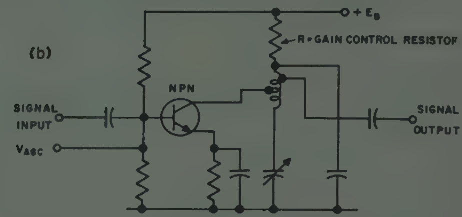


Fig. 15—Indirect V_c-control of tuned amplifier with AGC voltage applied to base.

whereas for V_c-control the gain decrease per decade of V_c is 12 to 15 db. The higher value applies to the common emitter circuit, due to the previously mentioned fact that $h_{21,e}$ is more sensitive to I_e and V_c variations than $h_{21,b}$.

The gain does not vary appreciably at values of I_e exceeding 500 μa and at values of V_c exceeding 1 v. This enables the design of amplifiers with "delayed" AGC. By selecting an appropriate "no signal" operating point, reasonable delay characteristics can be obtained using either gain control principle.

Although the characteristics discussed were those of a General Electric rate-grown transistor, other transistor types exhibit similar gain control properties.

III. Circuit Considerations

Methods of Gain Control

In practical AGC circuits the variation of I_e or V_c must be performed economically, with a minimum expenditure of control power. The following consideration illustrates the problem. Figs. 12 and 13 show that for adequate gain in the presence of small signals, I_e must be of the order of 500 μa and V_c at least 1 v. With increasing signal level, in the case of I_e-control, I_e must be decreased to 20 μa or less, whereas if V_c-control is used, V_c must be reduced to 30 mv or less. If I_e or V_c are controlled *directly*, the dc control power required to achieve this reduction of I_e or V_c is considerable.

The necessary dc power for direct control is not always available from detectors or other sources of AGC power and, consequently, in some cases (especially with a diode detector), if direct control of I_e or V_c is desired, an additional dc amplifier must be inserted in the feedback loop to deliver this control power (Fig. 14) unless detection is performed at very high level.

Control power can, however, be saved and the control power requirements on the detector (or any other control source) reduced by using the controlled transistor amplifier simultaneously as a dc amplifier of the control signal.

An example for the case of I_e-control is shown in Fig. 15(a). I_e is varied indirectly by applying an appropriate control potential to the base. The transistor shown is *n-p-n* and, consequently, with increasing signal level, a decreasing positive voltage is required as AGC signal. This will result in a decrease of I_e. (For a *p-n-p* transistor, a decreasing negative voltage is needed.) The transistor amplifies the dc control signal and moderate variations of the base current will result in appreciable variations of I_e.

A similar procedure can be applied in case of V_c-control (Fig. 15(b)). The AGC voltage is applied to the base.

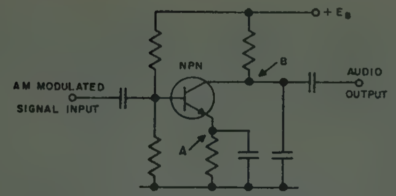


Fig. 16—Transistor detector circuit.

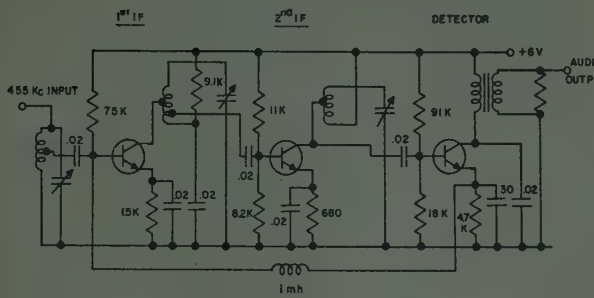


Fig. 17—Two stage IF amplifier followed by detector with V_c type AGC.

With increasing signal level, the AGC voltage acts to increase the emitter current I_e and the collector current I_c of the controlled stage. Due to the increased voltage drop developed by I_c across the resistor R inserted in the collector lead, V_c decreases and results in a reduction of gain. The controlled stage being *n-p-n*, an increasing positive control voltage is needed to reduce the gain (increasing negative for *p-n-p* transistor).

The necessary control power is often smaller in the case of V_c- than in the case of I_e-control, since very small variations of I_c will cause large variations of V_c, provided the gain control resistance R is sufficiently high.

Detector

The control voltage (with required polarity and sense of variation) can be obtained in many ways. Diode detectors produce positive or negative control voltages increasing with increasing signal level and by using appropriate biasing arrangements control voltages decreasing with increasing signal level can be easily produced.

In many cases (e.g., in broadcast receivers), transistor detectors delivering ample control power can be used. A possible arrangement is shown in Fig. 16. A positive voltage increasing with increasing signal level will appear at A. The potential at B will be decreasing with increasing signal level. Both points can be used as control signal sources. (With *p-n-p* detectors, the control voltage will be negative.)

AGC Circuits

Making use of the principles described in the previous paragraphs many, more or less different, AGC circuits can be designed. Blecher[1] and Barton[2] have described circuits based on I_e-control. Stern and Raper[3,4] use V_c-control in a broadcast receiver.

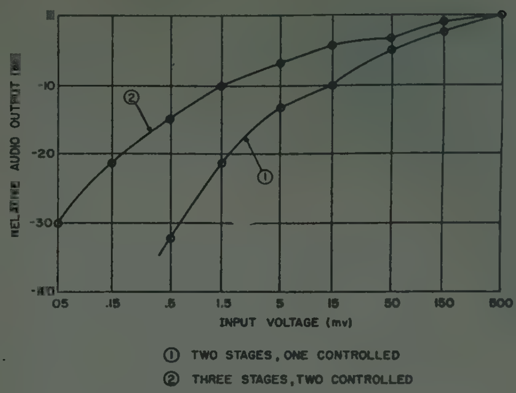


Fig. 18—AGC characteristics of IF amplifiers.

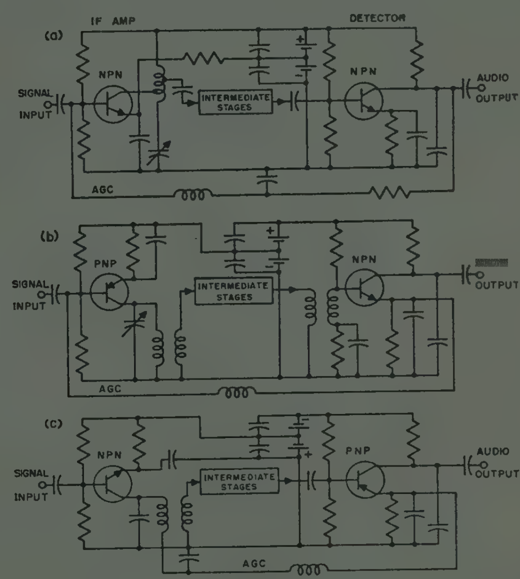


Fig. 19—AGC circuits.

A V_c-controlled circuit is shown in Fig. 17. The diagram represents a two-stage IF amplifier followed by a transistor detector. The first IF stage is V_c-controlled. The gain control characteristic of this amplifier is shown in Fig. 18.

Fig. 19 shows other possible circuits. Fig. 19(a) represents an I_e-controlled system, the performance of which is comparable to the one of Fig. 17. The circuit of Fig. 19(b) operates in a similar manner, but uses a *p-n-p* amplifier and a *n-p-n* detector. In the arrangement of Fig. 19(c), the collector voltage of the amplifier stage is controlled directly by the AGC voltage. Obviously, many other variations are possible.

The methods described can be applied to control the gain of several stages simultaneously. Desirable differential delays between the AGC action of different stages can be achieved, by either operating the controlled stages at different quiescent ("no signal") operating points or by designing different feedback networks for the controlled stages, or by deriving the control signal

for one controlled stage from another controlled stage.

Fig. 18 also shows the gain characteristic of a V_c-controlled IF amplifier two stages of which were controlled. The circuit was that of Fig. 17 with an additional controlled IF stage preceding the IF amplifier.

Distortion

Both principles of gain control (I_e and V_c) involve the reduction of the signal handling capability of the controlled stage at high signal levels (i.e., at reduced gain). I_e-control is achieved by decreasing I_e in presence of strong signals and, therefore, at small values of I_e, the permissible input current swing is reduced. If, on the other hand, the gain is decreased in the presence of strong signals by reducing V_c, the output voltage swing of the controlled stage is strongly limited.

In other words, the AGC performance of most transistors can be compared to that of sharp-cutoff vacuum tubes. Some transistors do, however, exhibit an α which starts to decrease at relatively large values of I_e or V_c. With such transistors, in common emitter configuration, the distortion problem is less serious.

In amplitude modulated systems, the controlled stages must be low-level stages to prevent distortion (or, even worse, the suppression) of the modulation envelope. (This does not apply, of course, to frequency-modulated systems.) In broadcast receivers, for instance, this limitation implies that AGC must not be applied to the last IF stage, and if more than one stage is controlled, suitable staggering of delays will be necessary. Controlled stages should be designed to handle only a fraction of a microwatt signal power. By careful design, adequate performance with tolerable distortion can be obtained. I_e-controlled stages can handle somewhat more signal power than V_c-controlled stages.

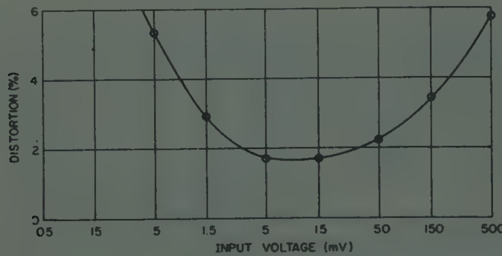


Fig. 20—Harmonic distortion as a function of input voltage.

Fig. 20 shows the distortion measured at the detector output of the amplifier of Fig. 17. High distortion at very low input signal levels is due to detector nonlinearity. The distortion is minimum at intermediate input levels and increases at high input levels as a result of AGC.

Bandwidth and Tuning

It has been seen that from the point of view of AGC the variation of h_{11} and h_{22}, and consequently that of input and output impedances is very important. The driving point impedances of a gain controlled stage vary with the signal level. This implies changes in the band-pass characteristics of tuned amplifiers:

1. Variation of the resistive component of the driving point impedances means variable damping of the interstage tuned circuits and consequently variation of bandwidth and selectivity.
2. Variation of the reactive component of the driving point impedances means detuning, i.e., a shift of the center frequency of the tuned amplifier.

In the case of emitter current control, resistive component of input impedance rises with decreasing I_e. Variation of the output impedance is less pronounced, but it also increases. If a parallel-parallel interstage tuning arrangement is used (Fig. 21(a)), with increasing signal level, that is with decreasing I_e, the bandwidth will decrease. The parallel-series tuning arrangement (Fig. 21(b)) is more desirable. The parallel output impedance of the first stage tends to decrease the bandwidth in the presence of large signals, but this is overcome by the increasing input impedance of the second stage which is in series with the tuned circuit. The combined effect is a moderate increase in bandwidth.

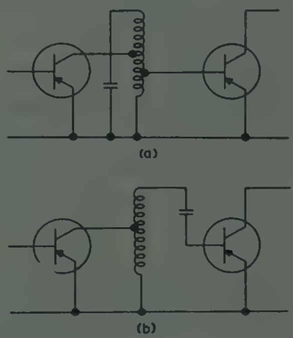


Fig. 21(a)—Parallel-parallel coupling circuit. (b) Parallel-series coupling circuit. (The circuits are drawn ac-wise.)

The situation is different in the case of collector voltage control. The input impedance is hardly at all affected by variations of V_c and the output impedance decreases if V_c is reduced. Consequently, the parallel arrangement leads to increasing bandwidth in the presence of strong signals. Fig. 22 (opposite page) shows the audio response of a V_c-controlled receiver as a function of the input signal. The increasing bandwidth at strong signal levels results in an "automatic tone control" feature, permitting taking advantage of the radiated spectrum of local transmitters.

In cases where it is desired to keep the bandwidth rigorously constant, it is necessary to place the desired selectivity into tuned circuits not adjoining controlled stages. Keeping the bandwidth of variable center frequency amplifiers constant is a particularly difficult

problem, and can often be achieved only by using stabilizing resistors and thereby sacrificing gain.

In the case of the I_e-controlled common emitter stage both input and output capacitances decrease in the presence of strong signals and result in an upward shift of the center frequency, The common base circuit usually shifts in the downward direction, because its input impedance is inductive and the inductive component increases with decreasing I_e.

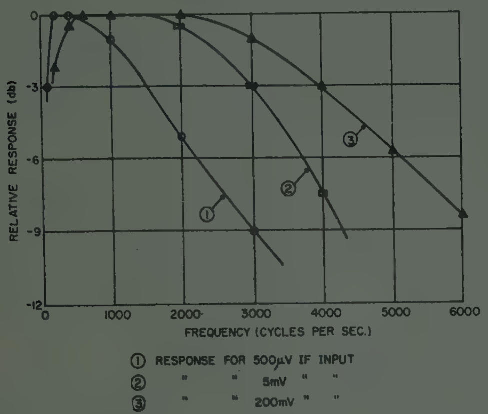


Fig. 22—Audio response of controlled IF amplifier-detector.

In the case of V_c-control the variation of the output capacitance is important; it increases with decreasing collector voltage and results in a downward shift of the center frequency.

An obvious method of center frequency stabilization is the use of a large external tuning capacitance, thereby reducing the effect of transistor reactances as tuning elements. If this is unsatisfactory, stabilizing resistances can be used; an uneconomical solution because of the loss in gain. In some cases it is advantageous to design the controlled stages as wideband amplifiers and to provide the selectivity by tuned circuits which are not associated with controlled stages.

Gain Control of Converters

Converter stages have special properties distinguishing them from linear amplifiers and special techniques can be used to control their gain.

Conventional I_e- or V_c-control will also be effective in the case of converters. The gain of converters[12] having a maximum at a certain value of I_e (Fig. 23), the gain can be reduced not only by decreasing I_e, but also by increasing it beyond the maximum. A further possibility of gain control exists due to the dependence of the converter gain on the injected oscillator voltage.

It is possible to perform with a single junction transistor both functions of oscillation and conversion simultaneously (Fig. 24). Such a circuit has optimum gain at

[12] J. Zawels, "The transistor as a mixer," Proc. IRE, vol. 42, pp. 542–548; March, 1954.

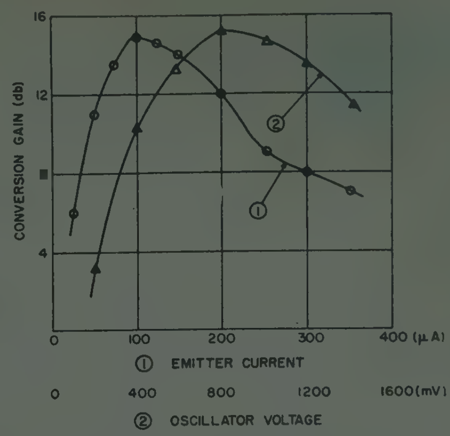


Fig. 23—Conversion gain as a function of emitter current and oscillator voltage.

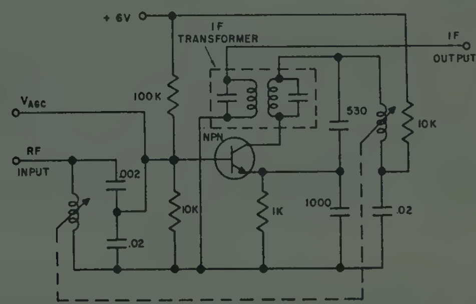


Fig. 24—Single transistor oscillator-converter with inductive tuning and AGC.

a certain value of I_e. Increasing or decreasing of I_e will result in reduction of gain and both methods can be used to obtain efficient automatic gain control.

IV. Conclusion

Automatic gain control of transistor amplifiers is possible due to the dependence of the gain on the dc operating point. AGC circuits with satisfactory performance have been built using either collector voltage or emitter current control. Efficient use can be made of the available control power if the controlled transistor is operated as a dc amplifier of the control signal.

No transistors are as yet available which have been designed specifically for AGC use. With present transistors, AGC involves problems of distortion, bandwidth and center frequency variation. In many applications these problems can be solved by appropriate design.

Acknowledgment

The authors are pleased to express their gratitude to J. A. A. Raper, who has designed part of the circuits and has done part of the experimental work leading to this paper. The helpful advice of R. L. Pritchard and the extensive measurements of C. D. Aiken are also gratefully acknowledged.

Two Network Theorems for Analytical Determination of Optimum-Response Physically Realizable Network Characteristics*

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Summary—By a technique combining variational methods, contour integration, and the concept of analytic extension, two network theorems are derived which form the basis of a procedure for determining the physically realizable network characteristic which is the best compromise between conflicting requirements in the sense of least mean-square error or some other criterion of approximation selected by the designer and written into the variational problem. Both the specifications and the criterion for compromise may include functions which are given as curves versus frequency rather than explicit functions of frequency.

The Wiener-Kolmogoroff theory of smoothing and prediction and the vestigial sideband filter with linear phase shift are included as illustrative application problems. For the former an alternative derivation in the frequency domain as well as a simpler procedure for numerical computations are given. For the latter a criterion for selecting the constant slope of the phase characteristic is derived in addition to providing the procedure for computing the optimum response filter.

I. Introduction

IN RECENT years, the primary emphasis of network synthesis appears to have been on the assembling of circuit elements to meet or approximate prescribed, physically-realizable attenuation and phase characteristic as functions of frequency. One may ask a rather basic question: "Are the prescribed characteristics best suited to the job for which the network is assigned?" If the answer to the above question is uncertain, then it would appear rather meaningless to try to fit the prescribed characteristics to one-tenth of a decibel or one degree. Many present-day problems contain conflicting requirements so that a network characteristic selected by qualitative judgment alone, even at its best, belongs to the uncertain category.

This paper describes an analytical method for obtaining a physically realizable network characteristic which is the best compromise towards meeting various desired but conflicting requirements. The "best" compromise may be based on a root-mean-square error criterion or other criterion of approximation as selected by the designer and introduced in the formulation of the minimization problem. The procedure is based on two network theorems which will be stated and proved in this paper.

One noteworthy point about this procedure is that while the derived network characteristic is physically realizable, it is in general not of the minimum phase type.[1] In other words, in selecting the best characteristic, the possibilities of using phase corrective networks are already taken into account. Thus the solution arrived at by this procedure gives the best final result including the phase corrective networks. One obvious byproduct of this procedure is that the proper phase corrective networks to be used are already prescribed in the specification of network characteristics and no laborious trial and error procedure is required.

Two examples are given to illustrate the mechanics of this method:

1. *Wiener's Prediction Network*[2,3]

The problem of predicting the instantaneous value with least mean-square error is shown to be analytically equivalent to designing for a nonrealizable network function which is simultaneously $e^{j\omega\tau}$ for the signal and zero for the noise. Wiener's prediction filter emerges as the best compromise.

Besides giving an alternative derivation, the present procedure also provides a simpler method for numerical computations of the prediction filter network function from arbitrary signal and noise power spectra. The new method does not require transforming back and forth between the frequency and time domains.

2. *Vestigial Sideband Filter*

In some cases of vestigial sideband transmission, notably television, the over-all (transmitter and receiver) filter response is required to have linear phase shift characteristic besides conforming to certain amplitude characteristic. (See Fig. 1, page 1130, for example.) It does not matter if the slope of the phase characteristic is larger or smaller. In terms of time delay, a constant time delay is not objectionable, but deviation from the constant time delay is to be minimized.

The above problem can be formulated into a weighted mean-square error minimization problem in which a designer may assign arbitrary weights to frequency ranges which he wishes to emphasize or de-emphasize. Such emphasis may be due to signal power spectrum or

* Original manuscript received by the IRE, March 16, 1955; revised manuscript received, June 6, 1955.

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[1] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., New York, N. Y., pp. 301–336; 1945.

[2] Norbert Wiener, "The Extrapolation, Interpolation and Smoothing of Stationary Time Series," John Wiley & Sons, New York, N. Y.; 1949.

[3] H. W. Bode and C. E. Shannon, "A simplified derivation of linear least square smoothing and prediction theory," Proc. IRE, vol. 38, pp. 417–425; April, 1950.

other considerations. The designer may select any slope for the phase characteristic. However, as a guide to such selection, it will be proven in general that the over-all error of the optimum filter decreases with increasing value of the assigned slope of the phase characteristic. This error reduction consideration is to be balanced against the increased complexity of the filters when a larger slope of the phase characteristic is used.

Once the minimization problem is formulated, a designer may follow a definite procedure to calculate the optimum response network function. An equation for the over-all error is also given which allows a designer to determine the error before the calculation of the optimum response network function is completed.

II. Definition of Terms

To describe the same terminal pairs of the same network, two entirely different network functions may be used. One may define the network function as the ratio of the input to the output, or as the logarithm of this ratio, or as the ratio of output to the input. The latter definition will be used throughout this paper. For passive as well as stable active linear networks, all the poles of the network function are in the left-half p-plane ($p=\sigma+j\omega$), excluding the imaginary axis. The above condition will be the basis for the analysis. While it may appear to be more restrictive than Bode's definition of physically realizable network functions, the restriction is due to a more definite definition rather than on the type of networks to be analyzed.

Some of the functions involved in the two theorems are defined only along the positive half of the real frequency axis. For instance, the desired phase and amplitude characteristics and the weighting factors may be given as curves versus frequency, rather than explicit expressions of frequency. As the required numerical computations of the design procedure being proposed in this paper are done entirely along the real frequency axis, specifications of the above type are entirely adequate for actual design work. However, in the theorems and derivations these functions are considered as defined over the entire complex plane and are analytic except at isolated singularities. The implied extension follows from a well-known mathematical theorem which states that if an analytic function is defined for the entire imaginary axis, then that function is completely determined over the complex plane. In theory at least, its value can be calculated for any arbitrary complex value of p by means of analytic extension.

Some explanatory remarks are due here on the choice of notations. In the literature the network function $G(j\omega)$ is usually expressed as a function of p or $j\omega$ along the real frequency axis while the spectral densities $S(\omega)$ and $N(\omega)$ are usually given as curves or real functions of ω. It appears desirable to the writer to comply with the conventional usage in both cases so that practically inclined readers will not find the design procedure confusing.

The independent variable $p=\sigma+j\omega$ is used outside the real frequency axis and the functions $G(j\omega)$, $S(\omega)$, and $N(\omega)$ etc., become $G(p)$, $S(-jp)$, and $N(-jp)$ etc., respectively. One should note that $S(-jp)$ and $N(-jp)$ etc., are actually even and real functions of p since $S(\omega)$ and $N(\omega)$ are even functions of ω.

Throughout LHP and RHP denote interior of the left-half p-plane and right-half p-plane, respectively.

In the following sections examples of application will be taken up first to illustrate the nature of the problems and the manner in which the two theorems will be applied. The derivations of the two theorems will follow the examples. This sequence is probably preferred by practically inclined readers. Theoretically inclined readers may find it more convenient to read Sections V and VI ahead of Sections III and IV.

III. The Prediction Problem in Presence of Noise

A filter is to be designed such that when a desired signal $s(t)$ plus a disturbing signal $n(t)$ is the input, the output $s_0(t)+n_0(t)$ is closest to $s(t+\tau)$ in the least mean-square error sense. The spectral densities of $s(t)$ and $n(t)$ are known and $s(t)$ and $n(t)$ are not correlated. The time interval τ is the forecast time.

The problem is simply to minimize the following integral, which represents the averaged value of the instantaneous error squared:

$$I=\lim_{T\to\infty}\frac{1}{2T}\int_{-T}^{T}[s_0(t)+n_0(t)-s(t+\tau)]^2dt. \quad (1)$$

Since $n(t)$ is not correlated with $s(t)$, $n_0(t)$ is not correlated with both $s_0(t)$ and $s(t+\tau)$, and (1) becomes:

$$I=\lim_{T\to\infty}\frac{1}{2T}\int_{-T}^{T}\{[s_0(t)-s(t+\tau)]^2+n_0(t)^2\}dt. \quad (2)$$

By a method similar to that used by Phillips,[4] (2) can be written in terms of frequency functions, as

$$I=\frac{1}{2\pi}\int_0^\infty\{|G(j\omega)-e^{j\omega\tau}|^2S(\omega)+|G(j\omega)|^2N(\omega)\}d\omega. \quad (3)$$

In (3), $G(j\omega)$ is the response function of the filter; $S(\omega)$ and $N(\omega)$ are spectral densities of the desired signal and disturbing signal respectively.

To apply the minimization theorem, the integrand may be written as:

$$F_0=\frac{S(\omega)}{2\pi}[G(j\omega)-e^{j\omega\tau}][G(-j\omega)-e^{-j\omega\tau}] + \frac{N(\omega)}{2\pi}G(j\omega)G(-j\omega). \quad (4)$$

Let $C(p)$ be defined as the following function:[5]

[4] H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 278–279; 1947.

[5] In general, $C(j\omega)$ is defined as the functional derivative $\partial F_0/\partial G(-j\omega)$.

$$C(p) = \frac{S(-jp)}{2\pi}\,[G(p) - e^{p\tau}] + \frac{N(-jp)}{2\pi}\,G(p). \quad (5)$$

The Minimization Theorem which will be derived later states that if a network function $G_m(p)$ minimizes the integral of (3), the corresponding $C(p)$ will not have any poles in the LHP including the imaginary axis, and vice versa. This condition can be used to determine $G_m(p)$ completely. Let $Y_+(p)$ and $Y_-(p)$ be defined such that $Y_+(p)\cdot Y_-(p) = S(-jp) + N(-jp)$ and all the poles and zeros of $Y_+(p)$ are in the LHP, and all the poles and zeros of $Y_-(p)$ are in the RHP.

Dividing (5) by $Y_-(p)$, there results

$$\frac{2\pi C(p)}{Y_-(p)} + \frac{S(-jp)e^{p\tau}}{Y_-(p)} = G_m(p)Y_+(p). \quad (6)$$

Let

$$\phi(p) \equiv \frac{S(-jp)e^{p\tau}}{Y_-(p)}.$$

In general, $\phi(p)$ has poles in both the LHP and RHP. It will be shown later that $\phi(p)$ can be expressed as the sum of two components $\phi_1(p)$ and $\phi_2(p)$ such that all the poles of $\phi_1(p)$ are in the LHP and all the poles of $\phi_2(p)$ are in the RHP. Since all the poles of $C(p)/Y_-(p)$ are in the RHP and all the poles of $G_m(p)\,Y_+(p)$ are in the LHP it follows from (6) that

$$G_m(p) = \frac{\phi_1(p)}{Y_+(p)}. \quad (7)$$

If $S(\omega)$ and $N(\omega)$ are given as explicit functions of ω, $\phi_1(p)$ can be determined by writing $\phi(p)$ into partial fraction form and retaining only the terms with poles in the LHP. However, in most actual cases, $S(\omega)$ and $N(\omega)$ are given as curves versus ω for real values of ω only. Then the gain and phase of the optimum response network function for real values of ω can be calculated as follows: Let $\alpha(\omega)$ and $\beta(\omega)$ be the gain and phase respectively of $Y_+(j\omega)$. As both $S(\omega)$ and $N(\omega)$ are positive and are even functions of ω, it follows that $-\beta(\omega)$ is the phase of $Y_-(J\omega)$ and that $|Y_+(j\omega)| = |Y_-(j\omega)| = \sqrt{S(\omega)+N(\omega)}$. The gain $\alpha(\omega)$ can be calculated as:

$$\alpha(\omega) = \tfrac{1}{2}\log_e\,[S(\omega) + N(\omega)]. \quad (8)$$

Since all the poles and zeros of $Y_+(j\omega)$ are in the LHP, $\beta(\omega)$ can be determined from $\alpha(\omega)$ by the well-known Bode's equations. Let $\beta(\omega_c)$ be the phase at an arbitrary frequency ω_c. It can be calculated as

$$\beta(\omega_c) = \frac{1}{\pi}\int_{-\infty}^{\infty} \frac{d\alpha}{du}\log\coth\frac{|u|}{2}\,du. \quad (9)$$

In (9) and subsequent equations, $u = \log(\omega/\omega_c)$. From (8) and (9) $\phi(j\omega)$ can be calculated. The real component of $\phi(j\omega)$ is

$$A(\omega) = \frac{S(\omega)}{\sqrt{S(\omega)+N(\omega)}}\cos\,[\omega\tau + \beta(\omega)]. \quad (10)$$

The imaginary component of $\phi(j\omega)$ is:

$$B(\omega) = \frac{S(\omega)}{\sqrt{S(\omega)+N(\omega)}}\sin\,[\omega\tau + \beta(\omega)]. \quad (11)$$

Presently one has only a set of numerical values of the real and imaginary components of $\phi(j\omega)$ instead of an explicit expression of $\phi(j\omega)$ in terms of ω. Apparently $\phi_1(j\omega)$ cannot be determined by the partial fraction method. It will be shown later in the Separation Theorem that the real and imaginary components of $\phi_1(j\omega)$ can be calculated as below: Let $A^1(\omega_c)$ and $B^1(\omega_c)$ for any arbitrary frequency ω_c be defined by the following equations.

$$A^1(\omega_c) = -\frac{1}{\pi\omega_c}\int_{-\infty}^{\infty}\frac{d[\omega B(\omega)]}{du}\log\coth\frac{|u|}{2}\,du \quad (12)$$

$$B^1(\omega_c) = \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{dA(\omega)}{du}\log\coth\frac{|u|}{2}\,du, \quad (13)$$

then:

$$\phi_1(j\omega) = \frac{1}{2}\,[A^1(\omega) + A(\omega)] + \frac{j}{2}\,[B^1(\omega) + B(\omega)]. \quad (14)$$

Let α_1 and β_1 be the gain and phase of $\phi_1(j\omega)$ in nepers and radians respectively as calculated from the values of $\phi_1(j\omega)$ obtained by (14). From (7) the gain and phase of $G_m(j\omega)$ are $\alpha_1 - \alpha$ and $\beta_1 - \beta$ respectively.

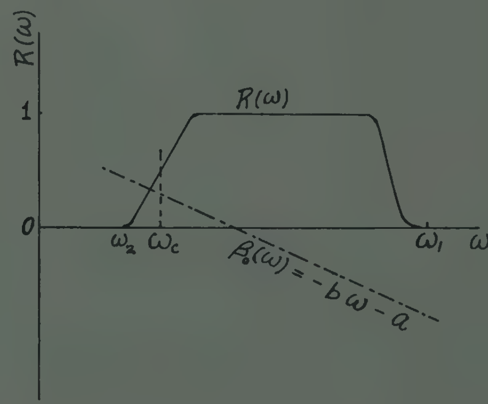

Fig. 1—Idealized response.

IV. The Vestigial Sideband Filter Problem

The vestigial sideband filter is designed for a given amplitude response characteristic and a linear phase characteristic as shown in Fig. 1. However, as this is not strictly realizable, the vector difference between the actual network response $G(j\omega)$ and the desired response is $G(j\omega) - R(\omega)e^{-j(b\omega+a)}$. From various engineering considerations, it is desirable to minimize a more general weighted square error integral

$$I = \int_0^{\infty} W(\omega)\,[G(j\omega) - R(\omega)e^{(-jb\omega+a)}]$$

$$\cdot\,[G(-j\omega) - R(\omega)e^{j(b\omega+a)}]\,d\omega \quad (15)$$

instead of a straight square error integral which will be obtained by setting $W(\omega)=1$. The function $W(\omega)$ will be called the spectral emphasis function. It is even and real and is determined by the following considerations: In the passband $W(\omega)$ may be the product of the signal spectral density and a weighting factor to take into account, for instance, that the picture detail may be considered more important than slow variations in luminous level in a television. While $W(\omega)$ at high frequencies is still smaller than $W(\omega)$ at low frequencies, it is not reduced to the same ratio as the spectral density. In the attenuation band, $W(\omega)$ determines the attenuation to a large extent. The larger $W(\omega)$ is, the larger will be the minimum value of the attenuation.

The optimum response network function $G_m(p)$ can be obtained by an analysis which is exactly parallel to that of the preceding section. From the Minimization Theorem,

$$G_m(p) = \frac{\phi_1(p)}{Y_+(p)}, \tag{16}$$

where $Y_+(p)Y_-(p)=W(-jp)$, and $\phi(p)=\phi_1(p)+\phi_2(p) = Y_+(p)f(p)$. All the poles and zeros of $Y_+(p)$ are in the LHP, and all the poles of $\phi_1(p)$ are in the LHP. The function $f(p)$ is the analytic extension of $f(j\omega)$ which is defined as follows:

$$f(j\omega) = R(\omega)e^{-j(b\omega+a)} \qquad \text{for } \omega > 0$$
$$f(j\omega) = R(\omega)e^{j(b|\omega|+a)} \qquad \text{for } \omega < 0$$

and either $R(\omega)$ is assumed to vanish at $\omega=0$ or $a=0$, so that there is no discontinuity at $\omega=0$. In case $R(\omega)$ is known only numerically, following the same reasoning as in the prediction filter case, a similar procedure for determining $G_m(j\omega)$ is obtained:

1. Calculate α and β as in (8) and (9), with the function $S(\omega)+N(\omega)$ replaced by $W(\omega)$.
2. Calculate $A(\omega)$ and $B(\omega)$ as follows:

$$A(\omega) = \sqrt{W(\omega)}R(\omega)\cos(\beta(\omega) - b\omega - a) \tag{17}$$
$$B(\omega) = \sqrt{W(\omega)}R(\omega)\sin(\beta(\omega) - b\omega - a). \tag{18}$$

The remaining steps are the same as (12), (13) and (14) etc., of the prediction filter case.

As $G_m(j\omega)$ is not expected to be appreciably affected by minor variations of $W(\omega)$, $W(\omega)$ may be written as a real algebraic function of ω^2, or $Y_+(j\omega)$ may be written as an algebraic function of $j\omega$ directly. The first step of the numerical computations will be saved in this case.

It is usually desirable to be able to calculate the minimum value of mean-square error before the optimum response network function is completely determined, and to know in advance what effect of change in the specified phase slope b would have on the minimum value of mean-square error. The following equations will be derived in Appendix I:

$$I_{\min} = \frac{1}{2}\int_0^\infty [A(\omega) - A^1(\omega)]^2 d\omega$$
$$= \frac{1}{2}\int_0^\infty [B^1(\omega) - B(\omega)]^2 d\omega \tag{19}$$

$$\frac{\partial I_{\min}}{\partial b} = -\frac{1}{\pi}\left[\int_0^\infty A(\omega)d\omega\right]^2$$
$$= -\frac{1}{\pi}\left[\int_0^\infty \sqrt{W(\omega)}R(\omega)\cos(\beta(\omega) - b\omega - a)d\omega\right]^2. \tag{20}$$

In (19) two alternatives for calculating $I_{\min}$ are given. $A^1(\omega)$ and $B^1(\omega)$ are calculated from (12) and (13). Since $B^1(\omega)$ is the simpler one to calculate, the latter form is recommended. Eq. (20) gives a valuable criterion for choosing b, as values of $A(\omega)$ can be readily calculated from the given functions. It means the following:

1. The approximation improves with increasing value of the phase slope b.
2. As a larger value of b means a more complicated filter construction, it is not desirable to make b unduly large. Eq. (20) predicts a case of "diminishing returns" as follows: With a small value of b, the argument of the cosine factor varies over a relatively small angle as ω varies over the range ω_2 to ω_1 for which $R(\omega)$ has substantial value. Therefore, $A(\omega)$ is of substantially the same sign and the magnitude of $\int_0^\infty A(\omega)d\omega$ is large. An increase in b tends to reduce $I_{\min}$ substantially. For a large value of b, the argument of the cosine factor increases many times 2π radians as ω increases from ω_2 to ω_1. Therefore, $A(\omega)$ has many sign reversals and the magnitude of $\int_0^\infty A(\omega)d\omega$ is small. A further increase in b will have very little effect in reducing $I_{\min}$. As a good compromise the value of b should be such that in addition to overcoming the upward phase slope of $Y_+(j\omega)$, adequate amount of downward slope is provided to cause a few sign reversals of $A(\omega)$, depending on the degree of approximation required.

V. Derivation of the Minimization Theorem for the Two Special Cases

It will be shown rigorously that for the two illustrative cases, the previous procedure leads to network functions for which the respective error integrals have absolute minimum values. For both cases the error integrals can be written as:

$$I = \int_0^\infty \{W_1(\omega)[G(j\omega) - f(j\omega)][G(-j\omega) - f(-j\omega)], + W_2(\omega)G(j\omega)G(-j\omega)\}d\omega \tag{21}$$

where $W_1(\omega)$ and $W_2(\omega)$ are positive real for real values of ω, and are even functions of ω. Let ω^{2n} and ω^{2m} be the leading terms of $W_1(\omega)$ and $W_2(\omega)$ respectively at the vicinity of $\omega\to\infty$, where n and m may be positive or negative integers.

For the prediction filter case, if n is 0 or positive, and $G(j\omega)$ is restricted to physically realizable functions, the

integral I does not converge at all. In other words, if the signal power does not diminish as frequency approaches infinity, no prediction is possible. If n is a minus integer, and k is the larger one (or the less negative one) of m and n, I converges if the leading term of $G(j\omega)$ is $(1/\omega)^{k+1}$ or higher power of $1/\omega$. I does not converge otherwise.

For the vestigial sideband filter case $W_2(\omega)=0$, and $f(j\omega)$ vanishes for ω larger than ω_1. The necessary and sufficient condition for I to converge is simply that the leading term of $G(j\omega)$ is $(1/\omega)^{n+1}$ or higher power of $1/\omega$ in the vicinity of $\omega\rightarrow\infty$. For both cases, only the network functions for which I is convergent need to be considered, and these are the admissible ones in the sense of condition c of the more general theorem. (See Appendix II.)

Let $G_m(j\omega)$ be the network function which minimizes I, and $G(j\omega)$ be any other admissible network function. The difference function $H(j\omega)$ is defined by the following equation:

$$G(j\omega) = G_m(j\omega) + H(j\omega). \tag{22}$$

Substituting (22) into (21) there results:

$$\begin{aligned} I=&\int_0^\infty \{W_1(\omega)\,|G_m(j\omega)-f(j\omega)|^2+W_2(\omega)\,|G_m(j\omega)|^2\}d\omega\\ &+\int_0^\infty (W_1(\omega)+W_2(\omega))\,|H(j\omega)|^2 d\omega\\ &+\int_0^\infty \{[W_1(\omega)+W_2(\omega)]G_m(j\omega)\\ &-W_1(\omega)f(j\omega)\}H(-j\omega)d\omega\\ &+\int_0^\infty \{[W_1(\omega)+W_2(\omega)]G_m(-j\omega)\\ &-W_1(\omega)f(-j\omega)\}H(j\omega)d\omega. \end{aligned}$$

Since the second integral is positive definite, it follows that the necessary and sufficient condition for $G_m(j\omega)$ to be the minimizing network function is that the sum of the last two integrals vanish. Combining these two terms, and using p as the independent variable, there results

$$\int_{-j\infty}^{j\infty} \{[W_1(-jp)+W_2(-jp)]G_m(p) - W_1(-jp)f(p)\}H(-p)dp = 0. \tag{23}$$

Since both $G_m(p)$ and $G(p)$ are of the order $(1/p)^{k+1}$ as $p\rightarrow\infty$, $H(p)$ is also of the order $(1/p)^{k+1}$ as a result of (22). The convergence at $\omega=\pm\infty$, and the vanishing of the integral along the large semicircle on LHP can be verified easily. The path of integration of (23) may be extended from $+j\infty$ through a large semicircle to $-j\infty$, enclosing the LHP, and there results:

$$\int_{\text{LHP}} \{[W_1(-jp)+W_2(-jp)]G_m(p) - W_1(-jp)f(p)\}H(-p)dp = 0. \tag{24}$$

Let $C(p)$ denote the expression inside the curved bracket in (24):

$$C(p)=[W_1(-jp)+W_2(-jp)]G_m(p)-W_1(-jp)f(p). \tag{25}$$

Since $H(p)$ does not have any poles in the RHP including the imaginary axis, it follows that $H(-p)$ does not have any poles in the LHP including the imaginary axis. A sufficient condition for satisfying (24) is that C does not have any poles in the LHP including the imaginary axis. This is also the necessary condition due to the arbitrariness of $H(-p)$. If $C(p)$ has poles in the LHP, a function $H(-p)$ can be found such that the sum of residues does not vanish, and the condition embodied in (24) will be violated.

VI. Separation Theorem

In carrying out the minimization calculations, usually there is required the separation of a function $\phi(j\omega)$ into two portions

$$\phi(j\omega) = \phi_1(j\omega) + \phi_2(j\omega), \tag{26}$$

such that all the poles of $\phi_1(p)$ are in the LHP, and all the poles of $\phi_2(p)$ are in the RHP. $\phi(p)$ does not have any poles on the imaginary axis.

In most problems $\phi(j\omega)$ is not expressed as an explicit function of ω, but with its amplitude and phase specified as curves versus ω for real positive values of ω. One well-known method of separation is by means of the Fourier transform. Let $\Psi(t)$ be the inverse transform of $\phi(j\omega)$. Generally $\Psi(t)$ does not vanish for $t<0$. Let $\Psi(t)$ be cut into two halves, $\Psi_1(t)$ and $\Psi_2(t)$, such that for $t<0$: $\Psi_1(t)=0$ and $\Psi_2(p)=\Psi(t)$; and for $t>0$: $\Psi_1(t)=\Psi(t)$ and $\Psi_2(t)=0$. Then $\phi_1(j\omega)$ and $\phi_2(j\omega)$ are obtained as Fourier transforms of $\Psi_1(t)$ and $\Psi_2(t)$. The disadvantage of this method is that $\Psi(t)$ is rather insensitive to $\phi(j\omega)$. Within reasonably expected accuracies of numerical computations, the result is not likely to be accurate enough for filter design work.

The separation theorem is an extension of Bode's relationships between real and imaginary components of physically realizable network functions. It may be stated as follows:

"If a function $\phi(j\omega)$ is bounded and is of the order of $1/\omega$ as ω approaches ∞, and

$$\phi(-j\omega) = \phi^*(j\omega) \tag{27}$$

where ϕ^* denotes the complex conjugate value of ϕ, then two functions $\phi_1(j\omega)$ and $\phi_2(j\omega)$ can be found such that $\phi(j\omega)=\phi_1(j\omega)+\phi_2(j\omega)$, and that both $\phi_1(j\omega)$ and $\phi_2(j\omega)$ satisfy the same conditions as $\phi(j\omega)$ for real values of ω, and that all the poles of $\phi_1(p)$ are in the LHP and all the poles of $\phi_2(p)$ are in the RHP $\phi_1(j\omega)$ and $\phi_2(j\omega)$ can be calculated as:

$$\phi_1(j\omega) = \tfrac{1}{2}[\phi(j\omega) + \phi^1(j\omega)] \tag{28}$$

$$\phi_2(j\omega) = \tfrac{1}{2}[\phi(j\omega) - \phi^1(j\omega)], \tag{29}$$

where the 'complementary function' $\phi^1(j\omega)$ is defined such that along the real frequency axis its real and

imaginary components $A^1(\omega)$ and $B^1(\omega)$ are related to the real and imaginary components $A(\omega)$ and $B(\omega)$ of the function $\phi(j\omega)$ by the following equations:

$$A^1(\omega_c) = -\frac{1}{\pi\omega_c}\int_{-\infty}^{\infty} \frac{d(\omega B)}{du} \log\coth\frac{|u|}{2} du \quad (30)$$

$$B^1(\omega_c) = \frac{1}{\pi}\int_{-\infty}^{\infty} \frac{dA}{du} \log\coth\frac{|u|}{2} du \quad (31)$$

where $u=\log_e (\omega/\omega_c)$."

The proof of the existence part of the theorem is simple. If $\phi_1(p)$ is defined as having the same poles and residues as $\phi(p)$ in the LHP, but none whatsoever in the RHP, and $\phi_2(p)$ is defined as having the same poles and residues as $\phi(p)$ in the RHP but none whatsoever in the LHP, then all the conditions of the first sentence of the theorem are satisfied.

To prove (28) to (31), let real functions $A_1(\omega)$, $A_2(\omega)$, $A^1(\omega)$, $B_1(\omega)$, $B_2(\omega)$ and $B^1(\omega)$ be defined by the following equations:

$$\phi_1(j\omega) = A_1(\omega) + jB_1(\omega) \quad (32)$$

$$\phi_2(j\omega) = A_2(\omega) - jB_2(\omega) \quad (33)$$

$$A^1(\omega) = A_1(\omega) - A_2(\omega) \quad (34)$$

$$B^1(\omega) = B_1(\omega) + B_2(\omega). \quad (35)$$

By definition, from (26), (32), and (33)

$$A(\omega) = A_1(\omega) + A_2(\omega) \quad (36)$$

$$B(\omega) = B_1(\omega) - B_2(\omega). \quad (37)$$

Solving for A_1, A_2, B_1 and B_2 from (34) to (37) leads to (28) and (29) directly. To prove (30) and (31), let the functions $\phi_3(j\omega)$ and $\phi_4(j\omega)$ be defined such that

$$\phi_3(j\omega) \equiv \phi_1(j\omega) + \phi_2(-j\omega) = A(\omega) + jB^1(\omega) \quad (38)$$

$$\phi_4(j\omega) \equiv \phi_1(j\omega) - \phi_2(-j\omega) = A^1(\omega) + jB(\omega). \quad (39)$$

The second equality signs in (38) and (39) follow from (32) to (37). As all the poles of $\phi_1(p)$ and $\phi_2(-p)$ are in the LHP, it follows that all the poles of $\phi_3(p)$ and $\phi_4(p)$ are in the LHP. By means of contour integration enclosing the RHP, one can easily prove that $A^1(\omega)$ and $B(\omega)$ are related by Bode equations, and that $A(\omega)$ and $B^1(\omega)$ are related by Bode equations. There is no limitation on the zero locations since $A(\omega)$ and $B^1(\omega)$ are real and imaginary components of the function $\phi_3(j\omega)$ itself rather than of the logarithm of $\phi_3(j\omega)$.

The following corollaries are easily verified:

1. "The complementary function of the complementary function is the function itself, provided that the function has no poles on the imaginary axis."
2. "If all the poles of a network function are in the LHP, excluding the imaginary axis, then the function is self-complementary."

Conclusion

A method for determining the optimum-response network function from conflicting requirements has been shown and illustrated by application to two actual examples: the prediction filter and the vestigial sideband filter with linear phase shift. The method is also applicable to cases where the desired performance is given as curves rather than explicit functions of ω, and the actual calculations involve evaluation of real integrals only along the real frequency axis.

The network theorems are proved rigorously for the special cases and are also stated in terms much more general than the two illustrative cases require, and can be expected to have wider applications (see Appendix II). Exact mathematical requirements such as continuity, etc., on the minimization function F has not been determined for the general case, but as most actual physical functions are continuous and have continuous derivatives, the problem rarely arises in application. However, such conditions as location of poles, convergence at infinity, etc., which are pertinent to engineering problems have been explicitly stated.

Appendix I

Derivation of Formulas Relating to the Minimum Value of Error for the Vestigal Sideband Filter Problem

Minimum Value of the Error Integral

From (16) and the definition of $\phi(p)$ one obtains:

$$G_m(j\omega) - f(j\omega) = \frac{\phi_1(i\omega)}{Y_+(j\omega)} - \frac{\phi(j\omega)}{Y_+(j\omega)} = -\frac{\phi_2(j\omega)}{Y_+(j\omega)}. \quad (40)$$

The complex conjugate of the above equation is:

$$G_m(-j\omega) - f(-j\omega) = -\frac{\phi_2(-i\omega)}{Y_+(-j\omega)} = -\frac{\phi_2(-j\omega)}{Y_-(j\omega)}. \quad (40a)$$

Substituting (40) and (40a) in (15), there results:

$$I_{\min} = \int_{\Pi}^{\infty} \phi_2(j\omega)\phi_2(-j\omega)d\omega. \quad (41)$$

From (18), (41) can be written as

$$I_{\min} = \int_0^{\infty} [A_2^2(\omega) + B_2^2(\omega)]d\omega. \quad (42)$$

Eq. (42) can be further simplified. Let the integral J be defined as:

$$J = \int_{-\infty}^{\infty} \phi_2^2(j\omega)d\omega. \quad (43)$$

Since $\phi_2(-j\omega) = \phi_2^*(j\omega)$, J can be written as

$$J = \int_0^{\infty} [\phi_2^2(j\omega) + \phi_2^{*2}(j\omega)]d\omega$$

$$= 2\int_0^{\infty} A_2^2(\omega)d\omega - 2\int_0^{\infty} B_2^2(\omega)d\omega. \quad (44)$$

As all the poles of $\phi_2(p)$ are in the RHP, and $\phi_2(p)$ is of the order of $1/p$ as p approaches infinity, it can be

easily shown that $J=0$ as a result of contour integration. Combining (42) and (44), there results

$$I_{\min} = 2\int_0^\infty A_2{}^2(\omega)d\omega = 2\int_0^\infty B_2{}^2(\omega)d\omega. \quad (45)$$

Effect of Time Delay or Slope of Phase Characteristics on Minimum Error

Differentiating (15), one obtains

$$\frac{\partial I}{\partial b} = 2\int_0^\infty W(\omega)\{R(\omega)e^{-j(\omega b+a)}[G(-j\omega) - R(\omega)e^{j(\omega b+a)}]$$
$$- R(\omega)e^{j(\omega b+a)}[G(j\omega) - R(\omega)e^{-j(\omega b+a)}]\}(j\omega)d\omega. \quad (46)$$

Substituting (40) and (40a) in (46), the latter becomes

$$\frac{\partial I_{\min}}{\partial b} = \int_0^\infty [\phi_1(-j\omega)\phi_2(j\omega) - \phi_1(j\omega)\phi_2(-j\omega)](j\omega)d\omega. \quad (47)$$

Eq. (47) can be rewritten into the following form:

$$\frac{\partial I_{\min}}{\partial b} = -\int_{-\infty}^\infty \phi_1(j\omega)\phi_2(-j\omega)(j\omega)d\omega. \quad (48)$$

The integrand of (48) is anlaytic and without singularities in the RHP. However, the integration over the large semicircular path does not vanish. In general, the amplitude of the desired response $R(\omega)$ is nonvanishing only in the limited range of ω. Therefore both $A(\omega)$ and $B(\omega)$ are nonvanishing in the same range only. From Bode's equation, one obtains, as $\omega_c \to \infty$.

$$B^1(\omega_c) = \frac{2\omega_c}{\pi}\int_0^\infty \frac{A}{\omega^2 - \omega_c{}^2}d\omega = \frac{B_\infty}{\omega_c} \quad (49)$$

$$A^1(\omega_c) = 0\left(\frac{1}{\omega_c{}^2}\right) \quad (50)$$

$$B_\infty = -\frac{2}{\pi}\int_0^\infty A(\omega)d\omega. \quad (51)$$

The real constant B_∞ is defined in (49). From (28) and (29), the limits of $\phi_1(j\omega)$ and $\phi_2(j\omega)$ are found to be $jB_\infty/2\omega$ and $-(jB_\infty/2\omega)$ respectively as ω approaches infinity along the real frequency axis. Since ϕ_1 and ϕ_2 are analytic, the results hold true for all large values of p. Therefore, by contour integration:

$$\int_{-\infty}^\infty \phi_1(j\omega)\phi_2(-j\omega)(j\omega)d\omega - \frac{\pi}{4}B_\infty{}^2 = 0. \quad (52)$$

From (48) and (52), one obtains (20).

Appendix II

General Formulation of the Minimization Theorem

While in most applications the minimization formulation is quite simple, it may be desirable to state the theorem in a relatively general form. It is:

If the integral along the real frequency axis (or the imaginary p-axis)

$$I = \int_0^\infty F_0(\omega, G_1, G_2 \cdots G_n, G_1{}^*, G_2{}^*, \cdots G_n{}^*)d\omega \quad (53)$$

is stationary with respect to infinitesimal variations of $G_1(p)$, $G_2(p)$ $\cdots$ etc., and correlated infinitesimal variations of $G_1{}^*$, $G_2{}^*$ $\cdots$ etc., under the following conditions, where $G_L{}^*$ is the shorthand notation for $G_L(-p)$: (1) the integrals along the real frequency axis

$$I_i = \int_0^\infty F_i(\omega, G_1, G_2 \cdots G_1{}^*, G_2{}^* \cdots)d\omega = K_i \quad (54)$$
$$i = 1, 2, 3, \cdots m$$

are fixed in value, that is, K_i's are constants; (2) $G_1(p)$, $G_2(p)$ $\cdots$ etc., are physically realizable network functions in the sense that they are real functions of p with all the poles in the left p-plane excluding the imaginary axis; (3) only sets of $G_1(p)$, $G_2(p)$ $\cdots$ etc. for which the integrals I_0, I_1 $\cdots$ converge and exist are admissible for comparison, and the functions F_0, F_1 $\cdots$ etc. are well behaved so that for the admissible $G_L(p)$'s differentiation under the integral sign is allowed;

(4) $$\left.\frac{\partial F_i}{\partial G_L}\right|_{p=j\omega_1} = \left.\frac{\partial F_i}{\partial G_L{}^*}\right|_{p=-j\omega_1}$$
$$i = 0, 1, 2, \cdots m; L = 1, 2, \cdots n \quad (55)$$

then the linear combinations of the functional derivatives

$$C_L(p) = \sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L{}^*} \quad (56)$$

do not have any pole in the LHP including the imaginary axis, where $\lambda_0 = 1$ and λ_1, λ_2, $\cdots$ λ_m are determined by K_1, K_2, $\cdots$ K_m or vice versa. The reverse statement is also true. That is, if $C_L(p)$ do not have any pole in the LHP inclusive, and given the conditions (1), (2), (3) and (4), then the integral I is stationary.

The proof of the theorem is as follows:

Let $EH_L(p)$'s be an arbitrary set of infinitesimal difference functions where E is an infinitesimal constant and $H_L(p)$'s are finite. Since the admissible $G_L(p)$'s must meet the condition of physical realizability, the difference functions $EH_L(p)$'s between two sets of admissible $G_L(p)$'s meet the same conditions. Making use of Lagrange multipliers, the first order variations are:

$$E\sum_{i=0}^m \lambda_i \int_0^\infty \sum_{L=1}^n \left[\frac{\partial F_i}{\partial G_L}H_L(p) + \frac{\partial F_i}{\partial G_L{}^*}H_L(-p)\right]d\omega = 0. \quad (57)$$

Since $H_L(j\omega)$'s are independent of each other, (57) is equivalent to

$$\int_0^\infty \sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L} H_L(p)d\omega + \int_0^\infty \sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L{}^*} H_L(-p)d\omega = 0 \quad (58)$$

$$L = 1, 2, \cdots n. \quad (58)$$

From (55):

$$\int_0^\infty \sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L} H_L(p)d\omega = \int_{-\infty}^0 \sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L{}^*} H_L(-p)d\omega. \quad (59)$$

Substituting (59) into (58), and using p as an independent variable, there results

$$\int_{-j\infty}^{j\infty} \left(\sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L{}^*} \right) H_L(-p)d(p) = 0. \quad (60)$$

As a result of condition (3), the integral on the left-hand side of (60) converges at $\omega = \infty$. Either the integrand is oscillatory and does not have a definite value at $\omega = \infty$, or the following equation must hold:

$$\lim_{p\to\infty} p \left(\sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L{}^*} \right) H_L(-p) = 0. \quad (61)$$

In the latter case, the path of integration of (8) may be extended from $+j\infty$ through a large semicircle to $-j\infty$, enclosing the entire LHP, and there results

$$\oint_{\text{LHP}} \left[\sum_{i=0}^m \lambda_i \frac{\partial F_i}{\partial G_L{}^*} \right] H_L(-p)d(p) = \oint_{\text{LHP}} C_L(p)H_L(-p)d(p) = 0 \quad (62)$$

$$L = 1, 2, \cdots n.$$

In a strictly physical problem, the functions F_i can always be formulated so that the integrand is nonoscillatory as the frequency approaches ∞, since the frequency range of interest is always limited. However, the integrand may become oscillatory in some over-simplifying mathematical formulations and in such cases, it would be necessary to check if the integration over the large arc does vanish before applying the minimization theorem.

Since $H_L(p)$ does not have any poles in the RHP including the imaginary axis, it follows that $H_L(p)$ does not have any poles in the LHP including the imaginary axis. A sufficient condition for satisfying (62) is that the functions $C_L(p)$'s do not have any poles in the LHP, including the imaginary axis. This is also the necessary condition due to the arbitrariness of $H_L(-p)$. If any of the functions $C_L(p)$ has poles in the LHP, a function $H_L(-p)$ can be found such that the sum of residues does not vanish, and the condition embodied in (62) will be violated.

Acknowledgment

The writer is indebted to Professor J. H. Mulligan, Jr., of the Department of Electrical Engineering of New York University for his many valuable suggestions.

Correspondence

VHF and UHF Signals in Central Canada*

Since the spring of 1953, the Radio Physics Laboratory of the Defence Research Board has carried out measurements on the amplitudes of vhf and uhf signals beyond the optical horizon. These signals have been propagated over rolling terrain between antennas whose height above ground was less than one hundred feet, but greater than one wavelength. Although the duration of the individual measurements was between one day and three weeks, the experiments were carried out over several different paths in central Canada during the winter and summer.

Fig. 1 shows the median signal power at 49, 91, 173 and 495 mc as measured at ranges between 20 and 235 miles. The reference axis (0 db) corresponds to the theoretical signal for diffraction around a smooth, spherical earth.[1] Here, the modified earth's radius factor was taken as 1.4 for summer propagation and 1.3 for winter propagation, according to Bean.[2] The signals at 21, 27 and 41 miles were measured over frozen muskeg in the sub-Arctic during January, while the signals at 30 and 49 miles were measured over Lake Ontario in October. The remaining signals at 29 and 50 miles and at greater ranges were measured between Ottawa and Sudbury. It will be seen that the diffraction theory predicts the median signal within a few decibels at ranges up to 30 or 40 miles. Beyond this range, the signals deviate widely from this theory. At ranges beyond 80 miles, the deviation from smooth-earth diffraction theory generally increases with increasing signal frequency. It should be mentioned that the measurements on different signal frequencies at ranges beyond 80 miles were made simultaneously at each location.

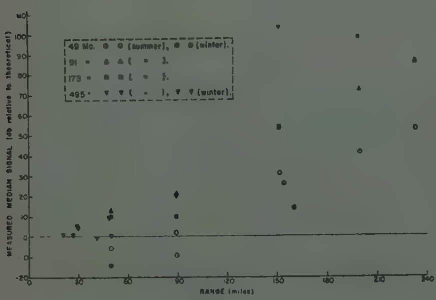

Fig. 1—Comparison between median measured signals and theoretical signals for diffraction around smooth spherical earth with atmosphere.

If it is assumed that the signals beyond 80 miles are propagated by scatter from atmospheric turbulence, the summer measurements indicate that the intensity of atmospheric turbulence varies inversely as the first power of height above ground. Fig. 2 shows the median of the random component of the signals at ranges beyond 80 miles, derived from the measured medians by removing the small steady diffraction component.[3] Examination of the data shows that the signal power (relative to free space level) varies approximately as the inverse cubed power of distance from the transmitter. According to the scatter formula recently proposed by Gordon,[4] this attenuation rate requires the distribution in atmospheric turbulence stated above. It is of interest to note also that the median measured signals lie well below the Bullington mean curve, which was fitted to radio measurements obtained during the past several years in the United States and the United Kingdom.[5]

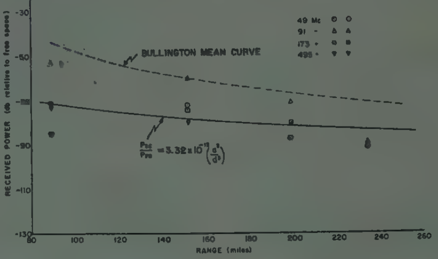

Fig. 2—Comparison between summer measurements and Gordon theory.

Further details on the measurements reported here are given in project reports of the Radio Physics Laboratory.

D. R. Hay and R. C. Langille
Radio Physics Lab.
Defence Research Board
Ottawa, Canada

* Received by the IRE, May 5, 1955.

[1] K. Bullington, "Radio propagation at frequencies above 30 megacycles," Proc. IRE, vol. 35, pp. 1122–1136; October, 1947.

[2] B. R. Bean, "The geographical and height distribution of the gradient of refractive index," Proc. IRE, vol. 41, pp. 549–550; April, 1953.

[3] A. P. Barsis, J. W. Herbstreit, and K. O. Hornberg, "Cheyenne Mountain tropospheric propagation experiments," Nat. Bureau of Standards Circular 554, p. 34; January 3, 1955.

[4] W. E. Gordon, "Radio scattering in the troposphere," Proc. IRE, vol. 43, pp. 23–28; January, 1955.

[5] K. Bullington, "Radio transmission beyond the horizon in the 40- to 4000 mc band," Proc. IRE, vol. 41, p. 135; January, 1953.

On Reciprocal Inductance*

Mr. Stockman's objection[1] to "inertiance"[2]—that most engineers would associate it with inductance rather than reciprocal inductance—is worthwhile, but answerable.

Clearly, with analogies based largely on arbitrary interpretation of coefficients and variables involved in basic differential equations, several models of analogous mechanical and electrical systems are possible and, in many circumstances, rather useful. Consequently, it is easy to make C, rather than $1/C$, correspond to the elastic constant of a spring, and so qualify for the name "elastance." In fact, the necessary argument would hinge upon electrical quantities that are the *duals* of the corresponding quantities in the argument that would make L correspond to inertia.

However, to arrive at analogies that are physically more satisfying, one would go back to the rudiments, and talk, not in terms of currents and voltages, but in terms of the more fundamental concepts of electric flux in a capacitor corresponding to the resting charge on it, and magnetic flux in a coil corresponding to the rate of moving charge through it. If the electric displacement is then associated with the length by which a spring is stretched, and the magnetic flux with the velocity of a moving mass, an elegant physical correspondence will be established between static and dynamic concepts. One consequence of this is that current and momentum become direct analogs—which does not lack in physical appeal. Also the analogies will be found to parallel electrically dual quantities. More important at the moment is the fact that $1/C$ will be found to deserve the name "elastance," and $1/L$, "inertiance."

Would Mr. Stockman prefer our "newtance" to his "yrneh" for the unit of inertiance—or even "newt"?

E. J. Baghdady
Res. Lab. of Elect.
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Cambridge, Mass.

* Received by the IRE, April 25, 1955.

[1] H. Stockman, "On reciprocal inductance," Proc. IRE, vol. 43, p. 341; March, 1955.

[2] E. J. Baghdady, "On reciprocal inductance," Proc. IRE, vol. 42, p. 1807; December, 1954.

Rebuttal[3]

Reading E. J. Baghdady's letter, I receive the impression that it serves as a derivation of his originally given formulas and statements, which are all correct. No statement to the contrary was implied in my correspondence item. As Baghdady points out, several models of analogous mechanical and electrical systems are possible, and one of these models will no doubt be selected as a future basis for terminology considerations. As stated, I feel most engineers would associate inertiance with inductance rather than reciprocal inductance. In view of the rapid progress of generalized field concepts, perhaps ten years from today most engineers would feel the opposite way!

It is not easy to formulate and define useful quantities and units, which are fundamentally correct, and one of the stumbling blocks is the fact that each generation inherits from the previous one a maze of nomenclature; to a great extent inadequate and erroneous. Mr. Baghdady deserves credit for having brought up for discussion in the correspondence section such fundamentals as the basic inductance concept, and perhaps the viewpoints he brings out illuminate the fact that many improvements would result, if other basic concepts were also brought out in open discussions. In fact, a correspondence page in every issue of the Proceedings of the IRE, reserved for new thoughts on symbols, terms, units, and definitions would, I am sure, be received with great interest by all readers, and would supply much fresh thought material for the various IRE symbols and standards committees. Actually, this sort of arrangement has already been successfully tried by two leading European scientific journals.

With regards to "yrneh," this term is not my "invention" any more. Somebody else invented it long before I did.[4] Probably, there have been many inventors since then.

Harry Stockman
Scientific Specialties Corp.
Boston 35, Mass.

[3] Received by the IRE, May 6, 1955.

[4] H. Stockman, "yrneh," Proc. IRE, vol. 43, p. 879; July, 1955.

Empirical Relationships with the Munsell Value Scale

In color television development and control activities, it is sometimes desirable to use or consider a gray scale which presents equal visual lightness steps. The Munsell value (renotation) scale is widely used for this purpose. Munsell value is defined as the lightness of the colors of a series of standards, varying in reflectance from zero to unity, in such a way that all visual intervals are uniform when viewed on a middle gray background by a light-adapted eye.[1]

The Weber-Fechner law has sometimes been applied to establish a visual scale linear with reflection density. However, it applies only to the dark-adapted eye; hence it is of limited practical utility. The Munsell value scale applies to the light-adapted eye; hence it is of more general use.

We have had occasion to consider scales of equal brightness differences for presentation on color picture tubes. A single visual scale can produce a scale of equal brightness differences on all color picture tubes only if the viewing conditions are standardized. "A visual scale is applicable only under the conditions of calibration or when the departures from these conditions have been found to make no difference."[2] The manner in which the conditions of observation influence brightness sensation is described and illustrated by Evans.[3]

It is granted that neither the Munsell value scale nor any other visual scale can be used "blindly" as a scale of equal brightness differences to be reproduced on color picture tubes, even if the color television system gives exact tone reproduction. Nevertheless, the Munsell value scale is an established equi-spaced visual scale for achromatic colors from black to white.

The utility of the Munsell value scale would be extended if there were a simple analytical relationship between Munsell renotation value (V) and per cent reflectance (R). This would facilitate conversion from physical to psychophysical units and vice versa. This relationship is now defined[4] by the empirical equation:

$$R = 1.2219\, V - 0.23111\, V^2 + 0.23951\, V^3 - 0.021009\, V^4 + 0.0008404\, V^5.$$

A more simple relationship between Munsell value and reflectance would be desirable.

In an attempt to find such a relationship, six equations were assumed, each of the form $V = a + b \cdot f(R)$. The empirical constants in each equation were found by the method of least squares. The goodness-of-fit for each equation is stated in parenthesis for the range of Munsell values from 1.0 to 9.0. These equations were calculated from the table of reference 4 using 0.10 value increments.

1. $V = -1.324 + 2.217\, R^{0.3520}$ (sigma$_v = \pm 0.018$)
2. $V = -1.636 + 2.468\, R^{1/3}$ (sigma$_v = \pm 0.029$)
3. $V = -3.598 + 4.146\, R^{1/4}$ (sigma$_v = \pm 0.12$)
4. $V = 0.336 + 1.010\, R^{1/2}$ (sigma$_v = \pm 0.17$)
5. $V = -0.388 + 4.505 \log_{10} R$ (sigma$_v = \pm 0.43$)
6. $V = 2.381 + 0.09834\, R$ (sigma$_v = \pm 0.66$).

For large ranges of value and for smaller ranges of value, these six empirical equation forms maintain approximately the same relative merit. The assumed linear relationship between reflectance and value (eq. 6) is very poor. The logarithmic relationship (eq. 5) is nearly as bad. The square-root and the fourth-root relationships (eqs. 3 and 4) are both somewhat better than the linear and logarithmic relationships. The cube-root of R (eq. 2) gives an excellent linear fit to value. The empirically-found 0.3520-power of R relationship (eq. 1) gives the best fit for a simple equation of the six forms tried.

In Fig. 1 we have compared the failure of the logarithmic equation with the failure of the simple cube-root of reflectance expression and the 0.3520-power of reflectance expression. A perfect fit would be represented by a horizontal line through the origin. The figure illustrates graphically the inadequacy of the logarithmic expression.

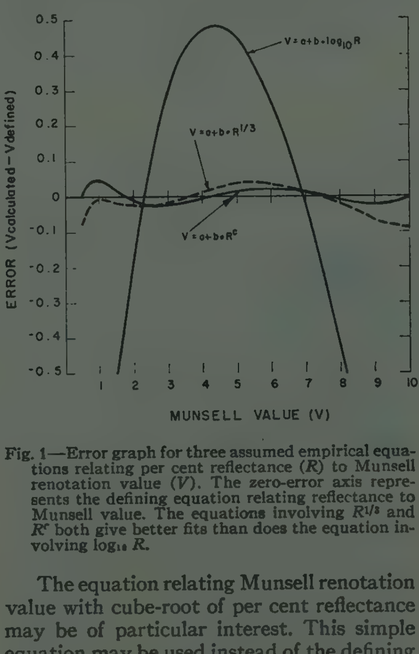

Fig. 1—Error graph for three assumed empirical equations relating per cent reflectance (R) to Munsell renotation value (V). The zero-error axis represents the defining equation relating reflectance to Munsell value. The equations involving $R^{1/3}$ and R^c both give better fits than does the equation involving $\log_{10} R$.

The equation relating Munsell renotation value with cube-root of per cent reflectance may be of particular interest. This simple equation may be used instead of the defining quintic equation for many purposes. The cube-root of R equation has the important advantage that it may be used to compute V from R or R from V. The defining quintic equation will not easily compute V from R.

J. H. Ladd and J. E. Pinney
Color Tech. Div., Eastman Kodak Co.
Rochester 4, New York

[1] A. E. O. Munsell, L. L. Sloan, and I. H. Godlove, "Neutral value scales. I. Munsell neutral value scale," *Jour. Opt. Soc. Amer.*, vol. 23, pp. 394–411; November, 1933.

[2] Committee on Colorimetry, Opt. Soc. Amer., "The Science of Color," Thomas Y. Crowell Co., New York, N. Y., p. 134; 1953.

[3] R. M. Evans, "An Introduction to Color," John Wiley and Sons, New York, N. Y., pp. 124–129, 157–172; 1948.

[4] S. M. Newhall, D. Nickerson, and D. B. Judd, "Final report of the O.S.A. subcommittee on the spacing of the Munsell colors," *Jour. Opt. Soc. Amer.*, vol. 33, pp. 385–418; July, 1943.

Effects of Impurities on Resonator Properties of Quartz*

During recent studies conducted at the Signal Corps Engineering Laboratories on fundamental properties of natural and synthetic quartz, it was found that piezoelectric resonators fabricated from synthetic quartz crystals[1] containing impurities introduced during the quartz growth cycle have different frequency-temperature characteristics from those made from natural quartz.

Measurements of frequency-temperature characteristics of fifth overtone, 29 mc AT-cut resonators prepared from aluminum-doped synthetic quartz show higher inflection temperature and orientation angles than those of natural quartz having similar frequency-temperature characteristics. This is graphically illustrated by the frequency-temperature curves of Fig. 1. In this figure

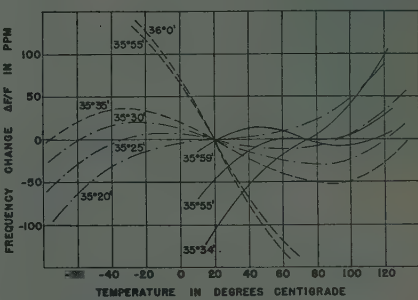

Fig. 1—Frequency temperature characteristics of AT-cut crystals with different ZZ' angles fabricated from natural and aluminum-doped synthetic quartz. The solid curves are the characteristics for aluminum-doped crystals; the dashed curves for natural quartz having same ZZ' angles; and dot-dashed curves for natural quartz having similar frequency temperature characteristics.

the solid line is used to represent aluminum-doped synthetic quartz resonators; the broken line represents the natural quartz resonators which have approximately the same orientation angles as the "doped" units; and the dot-dash line represents the natural quartz resonators which have the same frequency-temperature characteristics as those of the "doped" units. Angle given at ends of each curve is ZZ' orientation angle of particular AT-cut resonator.

The inflection temperature is defined as the temperature at which $d^2F/dT^2 = 0$. As Fig. 1 shows, the inflection temperature of the aluminum doped synthetic quartz resonators is at 75 degrees C., whereas that of natural quartz is at 20 degrees C. For similar frequency-temperature characteristics, the ZZ' orientation angle of aluminum doped synthetic quartz resonators is roughly 30 minutes higher than that of natural quartz.

The spectrochemical analysis of the aluminum-doped synthetic quartz shows that the following impurities (in parts per million by weight) are present: Al—100, Ge—50, Mg—10, Mn—10, Ca—5. Other impurity-doped synthetic quartz crystals have been grown and tested. In the case of germanium-doped quartz (Ge—1,000, Al—100, Fe—10, Ag—10, Mg—5, Ca—5), there is an upward shift of the inflection temperature of approximately 25 degrees, and an upward shift of the orientation angle of approximately 15 minutes. The investigation is being continued in an effort to determine the effects of various other impurities on resonator properties.

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Frequency Control Branch
Components Division
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* Received by the IRE May 23, 1955.

[1] A paper on synthesis of impurity doped quartz is in preparation by J. Stanley and S. Theokritoff.

Contributors

J. S. Ajioka (S'49) was born at Rexburg, Idaho, on August 9, 1923. He received his early college education at the University of California at Los Angeles. During World War II he served with the 442nd Infantry Regiment in southern Europe.

J. S. Ajioka

At the close of the war, Mr. Ajioka entered the University of Utah, where he received the B.S. degree in 1949 and the M.S. degree in 1951. Since graduation, he has been employed in the U. S. Navy Electronics Laboratory at San Diego, working on microwave antenna design and development, and in the Hughes Research Laboratories, Hughes Aircraft Co., Culver City, working on certain analytical phases of the radome design problem.

❖

F. R. Arams (S'44–A'49–SM'55) was born on October 18, 1925, in the Free City of Danzig. He received a B.S.E. in electrical engineering and a B.S.E. in mathematics from the University of Michigan in 1947, a M.S. degree in applied physics from Harvard University in 1948, and a M.S. degree in business management from Stevens Institute of Technology in 1953. He is presently studying for a doctorate in electrical engineering at the Polytechnic Institute of Brooklyn. During World War II he served as Communications Chief in charge of Radio Receiver Station WXH, Ketchikan, Alaska.

F. R. Arams

Mr. Arams joined the Radio Corporation of America in 1948 as a specialized trainee, and subsequently joined the Microwave Engineering laboratory of the RCA Tube Division, first at Lancaster, Pa. and now at Harrison, N. J. From 1948 to 1950, he was engaged in advanced development problems on microwave gaseous phenomena. From 1950 to 1953, he was project engineer on various magnetron and traveling-wave-tube design programs. From 1953 until early 1955, he served as assistant to the manager of Microwave Engineering, and at present he is in charge of Application Engineering on Microwave Tubes at RCA Tube Division.

Mr. Arams was a 1944 Donovan Scholar at the University of Michigan, and is a member of Eta Kappa Nu and Tau Beta Pi.

T. W. Butler, Jr. (A'53) was born in Niagara Falls, N. Y., on October 9, 1922. He received the B.S.E.E. and M.S. in physics from the University of Michigan in 1950 and 1953, respectively.

T. W. Butler, Jr.

From 1950 to 1951 he was employed as an assistant design engineer with Consumers Power Company, Jackson, Mich. From 1951 to the present time, Mr. Butler has been employed by the Engineering Research Institute, at the University of Michigan as an associate research engineer. He is currently engaged in both component and systems research programs and is working toward the Ph.D. degree in electrical engineering.

He is a member of Gamma Alpha and Sigma Xi.

❖

S. S. L. Chang (SM'53) was born in Peiping, China, in 1920. He received the M.S. in physics from Tsinghua University in China, in 1944 and the Ph.D. in electrical engineering from Purdue University in 1947. From 1947 to 1948, he taught at Purdue University.

S. S. L. Chang

He has been associated with Robbins and Myers, Inc., Springfield, Ohio since 1946. He joined the faculty of New York University in 1952 and is now Associate Professor of Electrical Engineering.

Dr. Chang is a member of the AIEE, ASEE, American Physical Society, Eta Kappa Nu, and Sigma Xi.

❖

For a photograph and biography of W. F. Chow, see page 881 of the July, 1955 issue of the Proceedings of the IRE.

❖

C. W. Helstrom was born on February 22, 1925, in Easton, Pa. From 1944 to 1946, he served as a radio technician in the United States Navy.

In 1947 he received the B.S. degree in engineering physics from Lehigh University, and in 1951 the Ph.D. in physics from the California Institute of Technology. Since 1951, he has been employed in the Electronics and Nuclear Physics Department of the Westinghouse Research Labs., in East Pittsburgh, Pa.

C. W. Helstrom

Dr. Helstrom is a member of the American Physical Society.

❖

H. K. Jenny (A'45–M'47–SM'50) was born on September 14, 1919, in Glarus, Switzerland. He received the M.S. degree in electrical engineering from the Swiss Federal Institute of Technology in Zurich, Switzerland in 1943. He remained there from 1943 to 1945, as assistant to Dr. F. Tank, head of the Institute of High Frequency.

H. K. Jenny

In 1946 he joined the Tube Division of the Radio Corporation of America in Lancaster, Pa., and has since been engaged in the development of microwave tubes. At present, Mr. Jenny is manager of the Microwave Tube Engineering Activity in Harrison, N. J.

❖

F. D. Lewis (S'36–A'38–VA'39–SM'50) was born in Liberty, Mo. on July 2, 1911. He received the A.B. degree at Central College, Fayette, Mo., in 1933, and the B.S. and M.S. degrees from Massachusetts Institute of Technology in 1937 and 1940. From 1937 to 1940 he was engaged in experimental work on uhf receivers and electromagnetic horn radiators at M.I.T. During the summer of 1940 he worked on Doppler-effect radar at Loomis Laboratory, subsequently going to the M.I.T. Radiation Laboratory when it opened in November, 1940.

F. D. Lewis

In 1941 he went to England as scientific liaison officer for the NDRC-OSRD. Returning in 1942, he became an expert con-

sultant in the Office of the Secretary of War on radar countermeasures and allied problems. He has been with the General Radio Co. since 1945, and has been working on frequency measurement since 1949.

Mr. Lewis holds the President's Certificate of Merit. He is a member of Sigma Xi and the AAAS.

❖

W. J. Lindsay (A'55) was born in Miles, Texas, on October 28, 1925. He received his B.S. degree from Texas A. and M. College in 1948 and the M.S. degree in 1954. From August, 1948, to May, 1951, he was employed by the Humble Oil and Refining Company in Houston, Texas as an assistant engineer. From May, 1951 to August, 1952, Mr. Lindsay was employed in the EWS Research Laboratory of the Halliburton Oil Well Cementing Company of Houston as project engineer. From 1952 to 1954 he was engineer in charge of a network analyzer project, at the Texas A. and M. Research Foundation.

W. J. Lindsay

Mr. Lindsay joined the staff of the Electronic Defense Group, Engineering Research Institute, at the University of Michigan in August, 1954 as research associate. He is currently working toward the Ph.D. degree in electrical engineering.

He is a member of Tau Beta Pi.

❖

L. W. Orr (A'52) was born in 1915 in Hamilton, Canada. He received the B.A.Sc. degree in electrical engineering in 1943 from the University of Toronto, and the M.S. and Ph.D. degrees in electrical engineering in 1946 and 1949 from the University of Michigan. During World War II, he was an officer in the Signals Branch of the Royal Canadian Air Force.

L. W. Orr

From 1950 to 1951, Dr. Orr was in charge of the research and development of magnetic devices at the Burroughs Research Division, Philadelphia, Pa. From 1949 to 1950, and from 1951 to the present time, he has been a research engineer in the Engineering Research Institute, at the University of Michigan. He is presently engaged in a component research program in the Electronic Defense Group.

For a photograph and biography of R. L. Pritchard, see page 105 of the January, 1955 issue of the Proceedings of the IRE.

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For a photograph and biography of A. P. Stern, see page 882 of the July, 1955 issue of the Proceedings of the IRE.

❖

D. Weighton was born on June 26, 1913, in Kingston-upon-Hull, England. He received the M.A. degree from Cambridge University in 1935. From 1936 to 1938, he served as Education Officer in the Royal Air Force.

D. Weighton

In 1938 Mr. Weighton joined the research staff of Pye Limited, Cambridge, Eng. and from 1948 has headed the research division of that company.

Mr. Weighton is an associate member of the British Institution of Electrical Engineers.

IRE News and Radio Notes

PG on Broadcast Transmission Systems Will Hold Symposium

"New Perspectives in the Field of Broadcasting" will be the theme of the Fifth Annual Fall Symposium of the Professional Group on Broadcast Transmission Systems. The meeting is scheduled for September 23 and 24 at the Hamilton Hotel in Washington, D. C. Registration will open at 9:00 a.m., September 23. Advance registration and reservations may be sent to PGBTS, Seventh Floor, 1735 DeSales Street, N.W., Washington 6, D. C.

The Friday morning session of the symposium will be devoted to 32 new television broadcasting equipment and facilities. Papers will include *Multiple Antenna System with Antennas of Equal Height* by L. J. Wolf of RCA, *Studio Switching Problems with Color Signals* by H. W. Morse of General Electric, and *A Fifty Kilowatt Transmitter* by John Ruston of DuMont. Philip B. Laeser, Technical Director of WTMJ in Milwaukee, will speak on *Integrating Color Equipment with Monochrome Facilities at your Television Station.*

The Friday afternoon session will be devoted to measurements, including propagation factors, in television. A paper, made available by Edward W. Allen, Jr., of the FCC and including work directed by William C. Boese and Harry Fine, *Present Knowledge of Propagation in VHF and UHF Television Bands*, will be delivered. Edward W. Chapin, Chief of the Laurel, Maryland Laboratory Division of the Federal Communications Commission will deliver a paper summarizing field intensity measurements which have been made on various UHF stations.

The Saturday morning session will include papers on new broadcast operation techniques and equipment arrangements. R. A. Isberg, of the Ampex Corporation, will deliver a paper on *Using New Tape and Film Techniques to Increase Broadcast Operation Efficiency.* A. C. Goodnow, Westinghouse Broadcasting Company, will speak on *Experimental Experience with Remote Control of High Power and Directional Antenna Broadcast Transmitter Operations. A Novel Television Slide Sequencing Arrangement* will be presented by Roger E. Peterson of WNBF, Binghamton, New York. Edgar F. Vandivere, Jr. of Vandivere, Cohen and Wearn will present a paper on *Some Techniques in Automatic Programming.*

Moderators for the sessions will be Ralph N. Harmon, Westinghouse Broadcasting Company; Stuart L. Bailey, Jansky and Bailey, Incorporated, and George C. Davis, consulting engineer.

A Friday evening cocktail hour will begin at 6:00 p.m. and will be followed by a banquet. Arrangements have also been made with Edgar T. Martin, Chief Engineer of the Voice of America for conducted tours there on Saturday afternoon.

Steering Committee includes Oscar Reed, Jr., General Chairman; Ralph N. Harmon, Technical Program Committee; Harold Dorschug, Public Relations and Publicity; C. M. Brown, Finance; Lewis Winner, Papers Review; and Irma B. Galane, Local Arrangements Committee.

W. C. White Contributes to Browder Thompson Fund

Friends of W. C. White recently presented him with several gifts upon his retirement from the General Electric Company. Mr. White has generously donated those gifts that were in cash to the Browder J. Thompson Fund. The money is to be used in six annual installments to augment the Browder J. Thompson Memorial Prize, which is given annually to an author, less than 30 years old, who has published the most outstanding paper in an IRE publication.

Mr. White has been active in IRE affairs for more than twenty-five years, serving as Treasurer in 1946 and Director from 1943 to 1947.

Calendar of Coming Events

IRE-ISA Tenth Annual Instrument Conference, Shrine Auditorium, Los Angeles, Calif., Sept. 12–16

Association for Computing Machinery, Annual Meeting, Moore School of Electrical Engineering, U. of Pa., Sept. 14–16

IRE Professional Group on Nuclear Science—Second Annual Meeting, Oak Ridge National Labs., Oak Ridge, Tenn., Sept. 14–17

IRE Cedar Rapids Section Symposium on Automation, Cedar Rapids, Ia., Sept. 17

PGBTS Fifth Annual Fall Symposium, Hamilton Hotel, Washington, D. C., Sept. 23–24

Symposium on Physiologic and Pathologic Effects of Microwaves, Mayo Clinic, Rochester, Minn., Sept. 23–24

RETMA Automation Symposium, U. of Pennsylvania, Philadelphia, Pa., Sept. 26–27

PG on Vehicular Communications, Sixth Annual Meeting, Multnomah Hotel, Portland, Ore., Sept. 26–27

IMSA Annual Convention, Hotel Seneca, Rochester, N. Y., Sept. 26–29

International Analogy Computation Meeting, Société Belge des Ingenieurs des Télécommunications et d'Electronique, Brussels, Belgium, Sept. 27–Oct. 1.

IRE-AIEE Conference on Industrial Electronics, Rackham Memorial Building, Detroit, Michigan, Sept. 28–29

National Electronics Conference, Hotel Sherman, Chicago, Ill., October 3–5

Audio Engineering Society Convention, Hotel New Yorker, New York City, Oct. 12–15

IRE-RETMA Radio Fall Meeting, Hotel Syracuse, Syracuse, N. Y., Oct. 17–19

Conference On Electrical Insulation, Pocono, Pa., Oct. 17–19

Eighth Annual Gaseous Electronics Conference, General Electric Res. Lab., Schenectady, N. Y., Oct. 20–22

PG on Electron Devices Annual Technical Meeting, Shoreham Hotel, Washington, D. C., Oct. 24–25

GAMM and NTG-VDE International Conference on Electronic Digital Computers, and Data Processing, Darmstadt, Germany, Oct. 25–27

IRE East Coast Conference on Aeronautical and Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md., Oct. 31–Nov. 1.

Symposium on Applied Solar Energy, Westward Ho Hotel, Phoenix, Ariz., Nov. 1–5

Kansas City Section Electronics Conference, Kansas City, Kansas, Nov. 3–4

IRE-AIEE-ACM Eastern Joint Computer Conference, Hotel Statler, Boston, Nov. 7–9

IRE-AIEE-ISA Electrical Techniques in Medicine and Biology, Shoreham Hotel, Washington, D. C., Nov. 14–16

IRE-PGCS Symposium on Aeronautical Communications—Civil and Military; Utica, New York, Nov. 21–22

PGI and Atlanta Section Data Processing Symposium, Hotel Biltmore, Atlanta, Ga., Nov. 28–30

Symposium on Reliability and Quality Control

Victor Wouk (left), Chairman of PG on Reliability and Quality Control, is Publicity Chairman for Symposium.

The PG on Reliability and Quality Control will sponsor a Symposium on Reliability and Quality Control in Electronics. The symposium will be held January 9 and 10, 1956 at the Hotel Statler in Washington.

A preliminary schedule for the symposium has been published.

First Session: "Quality Control and Automation," "Reliability of Complex Commercial Equipment," "Reliability of Weapons Systems."

Second Session: "The Control Chart Applied to Field Failure Data," "An Approach to the Study of System Reliability," "The Relation of Life Tests to Failure Rates." The panel discussions will include: "High Reliability vs Cost of Equipment and Parts," "System Testing vs In-process Control," "The Problem of Field Failure Reporting (including new Air Force plans)."

Third Session: "Advances in Tube Reliability," "Reliable Connectors Through Quality Control," "Controlling Relay Characteristics," "Reliable Capacitors," "Parts Reliability Research by the Services."

Fourth Session: "Quality Requirements and Acceptance Procedures."

Symposium on Microwave Techniques to Meet in February in Philadelphia

Jointly sponsored by the PG on Microwave Theory and Techniques, the PG on Antennas and Propagation, and the Philadelphia Section, a national symposium on Microwave Techniques will be held at the University of Pennsylvania in Philadelphia, February 2 and 3, 1956.

Sessions are planned to include the following topics:

Radiating Systems: radomes, techniques of antenna gain and pattern measurement, the use of near field measurements, problems associated with paraboloid antennas of 20 feet or more diameter.

Panel Session on Guided Microwave Transmission: rectangular guide, ridge guide, round guide, dielectric guide, single wire guide.

Components: filters, converters, duplexers, directional couplers, non-reciprocal components.

Propagation: scatter.

Measurements: power spurious emission, spurious modulation, emission bandwidth, gain standards.

Those interested in presenting papers on one of these subjects should submit 250 word abstracts no later than October 15. The abstracts, in duplicate, may be sent to D. R. Crosby, RCA, Bldg. 10-1, Camden 2, N.J.

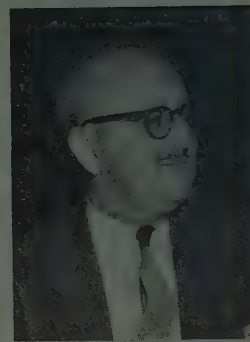
November 4 IRE Convention Papers Deadline

November 4 is the deadline for submission of 1956 IRE National Convention papers. Information needed, to be included in triplicate: a 100-word abstract with title of paper, name, and address; a 500-word summary, with title of paper, name, and address. One of the following fields in which the paper falls should be indicated. Aeronautical and Navigational Electronics, Antennas and Propagation, Audio, Automatic Control, Broadcast and Television Receivers, Broadcast Transmission Systems, Circuit Theory, Communications Systems, Component Parts, Electron Devices, Electronic Computers, Engineering Management, Industrial Electronics, Information Theory, Instrumentation, Medical Electronics, Microwave Theory and Techniques, Nuclear Science, Production Techniques, Reliability and Quality Control, Telemetry and Remote Control, Ultrasonics Engineering, Vehicular Communications.

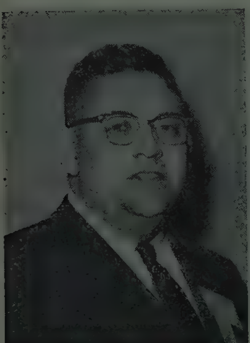
Send material to Russell R. Law, 1956 Technical Program Committee, IRE, 1 East 79 St., N. Y. 21, N. Y.

ANNUAL MEETING OF THE DALLAS-FORT WORTH SECTION

Seen at right are two participants at the Dallas-Fort Worth Section Annual Meeting. Bernard Brister (*left*), public relations counselor for the Section, was honored with membership in the Institute of Radio Engineers. Making the presentation is Dr. Gordon K. Teal of Texas Instruments, Inc., new Dallas-Fort Worth Vice-Chairman.



(*Above*) Mark Bullock, Continental Electronics Manufacturing Co., is caught in an informal shot during the banquet which was held in conjunction with the meeting. He will be the new chairman of the Section.



Shown above is John A. Green, outgoing Chairman of the Section. He was honored at the meeting for the services to the IRE during the past year.

Shown at left are the new contributing editors who were elected recently at the Annual Meeting. (*From left to right*) Kenneth P. Dowell, of the Texas Instruments, Inc., Dallas, and James M. Weldon of the Collins Radio Company, Cedar Rapids, Iowa, look over an issue of *Direction* with T. A. Wright, Jr., of A. Earl Cullum, Jr., and Associates, Dallas, John K. Godbey, and Theil Sharp, both of Collins Radio Company.

February to be Date for Western Computer Conference

The Western Computer Conference will be held in San Francisco February 8–10, 1956. It will be sponsored jointly by the IRE, AIEE, and ACM.

Papers on all phases of the computer field are now being solicited. In addition to paper title, authors are asked to submit an abstract of approximately 200 words, suitable for reproduction in the program, and either the complete manuscript or sufficient additional information to permit evaluation by the Technical Program Committee. Early submission of papers is desired, the final deadline being November 15. This is the latest date that is feasible and papers received thereafter cannot be considered.

Authors should indicate any plans for publication and should state what facilities, such as slide or movie projectors, power sources, etc. are required. For uniformity of handling, it is requested that all papers be directed to: Byron J. Bennett, Chairman, Technical Program Committee, Stanford Research Institute, Menlo Park, California.

Several Proceedings Available

The *1954 Eastern Joint Computer Conference Proceedings* are now available at $3.00 per copy. These are the complete papers that were presented in Philadelphia December 8–10, 1954. Also available at $3.50 per copy are the *1955 National Telemetering Conference Proceedings*. These represent the complete papers delivered at Chicago May 18–20. Both volumes may be obtained through IRE Headquarters, 1 E. 79 St., N. Y. 21, N. Y.

The *Proceedings of the 1955 Electronic Components Conference* held in Los Angeles are now available. The publication contains more than forty-eight papers on electronics which will not be duplicated in other publications. Checks for $4.50 per copy should be made payable to the 1955 Electronic Components Conference, 8820 Bellanca, Los Angeles, California.

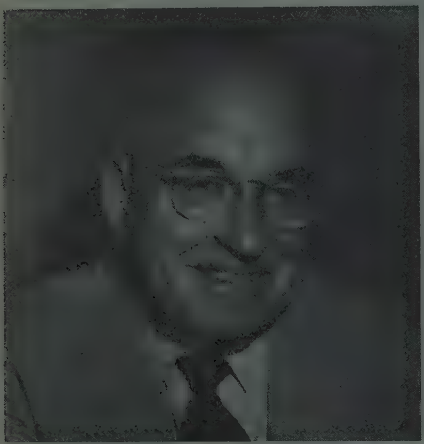
National Simulation Conference Will Meet in Dallas

The Dallas-Fort Worth Chapter of the PG on Electronic Computers will sponsor a National Simulation Conference in Dallas, Texas, January 19–21, 1956. The conference will be devoted to simulation and associated computing techniques, and will include topics in general simulation (mathematical, physical, logistic, etc.); advances in computer design, techniques, and applications; and methods of determining and improving the accuracy of analog solutions.

Papers for the conference are now being solicited. Although it is expected that most of the papers will deal with analog computers, papers on the use of digital computers in simulation will be strongly encouraged. Prospective authors should submit in duplicate by September 10, a 100 word abstract together with either a 500 word summary or the complete paper itself to: J. R. Forester, 2104 Huntington, Arlington, Texas.

Former IRE President Elected Head of Radio Pioneers Club

Raymond F. Guy recently has been elected president of the Radio Pioneers Club. He is Director of Radio Frequency Engineering for the National Broadcasting Company. Mr. Guy was on the original staff, composed of only a few persons, at WJZ when it was opened by the Westinghouse Company in Newark, N. J., in 1921 as the world's second broadcasting station. At that time the audience consisted of only a few amateurs. Commercial broadcasting was unknown and practically all operating methods and techniques had to be originated by trial and error. In the last 36 years, Mr. Guy has played a part in developing network broadcasting, short-wave broadcasting to foreign countries, frequency modulation and the evolution and development of television.



Raymond F. Guy, new President of Radio Pioneers

During World War II, he participated in projects of the Office of Strategic Services, the Co-ordinator of Inter-American Affairs, and the Office of War Information, one of which took him abroad. Since the war he has participated in international radio conferences in Havana, Mexico City, and Montreal.

A fellow of the Radio Club of America, a charter member of the Radio Pioneers Club, a life member of the Veteran Wireless Operators Association, and a member of the Society of Professional Engineers, he was admitted to practice as a professional engineer in New York and New Jersey in 1937. He has also served on the Radio Technical Planning Board and represented the IRE in the activities of the American Standards Association.

Mr. Guy became an IRE Associate Member in 1925, a Member in 1931 and a Fellow in 1939. He was President in 1950. In 1944, he was elected to the Board of Directors and served through 1948, including one term as Treasurer. He has served on a number of IRE committees, functioning as chairman of the standards, public relations, founders, transmitters, membership and office practices committees, and vice chairman of the building fund and executive committees.

TRANSACTIONS OF THE IRE PROFESSIONAL GROUPS

The following issues of TRANSACTIONS are available from The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., at the prices listed below:

Sponsoring Group	Publications	Group Members	IRE Members	Non-* Members
Aeronautical & Navigational Electronics	PGAE-5: A dynamic Aircraft Simulator for Study of Human Response Characteristics (6 pages)	$ .30	$ .45	$ .90
	PGAE-6: Ground-to-Air Cochannel Interference at 2900 NC (10 pages)	.30	.45	.90
	PGAE-8: June 1953 (23 pages)	.65	.95	1.95
	PGAE-9: September 1953 (27 pages)	.70	1.05	2.10
	Vol. ANE-1, No. 1, March 1954 (51 pages)	1.00	1.50	3.00
	Vol. ANE-1, No. 2, June 1954 (22 pages)	.95	1.40	2.85
	Vol. ANE-1, No. 3, September 1954 (27 pages)	1.00	1.50	3.00
	Vol. ANE-1, No. 4, December 1954 (27 pages)	1.00	1.50	3.00
	Vol. ANE-2, No. 1, March 1955 (41 pages)	1.40	2.10	4.20
Antennas and Propagation	PGAP-4: IRE Western Convention, August 1952 (136 pages)	2.20	3.30	6.60
	Vol. AP-1, No. 1, July 1953 (30 pages)	1.20	1.80	3.60
	Vol. AP-1, No. 2, October 1953 (31 pages)	1.20	1.80	3.60
	Vol. AP-2, No. 1, January 1954 (39 pages)	1.35	2.00	4.05
	Vol. AP-2, No. 2, April 1954 (41 pages)	2.00	3.00	6.00
	Vol. AP-2, No. 3, July 1954 (36 pages)	1.50	2.25	4.50
	Vol. AP-3, No. 4, October 1954 (36 pages)	1.50	2.25	4.50
	Vol. AP-3, No. 1, January 1955 (43 pages)	1.60	2.40	4.80
	Vol. AP-3, No. 2, April 1955 (47 pages)	1.60	2.40	4.80
Audio	PGA-7: Editorials, Technical Papers & News, May 1952 (47 pages)	.90	1.35	2.70
	PGA-10: November–December 1952 (27 pages)	.70	1.05	2.10
	Vol. AU-1, No. 1, January–February 1953 (24 pages)	.60	.90	1.80
	Vol. AU-1, No. 2, March–April 1953 (34 pages)	.80	1.20	2.40
	Vol. AU-1, No. 3, May–June 1953 (34 pages)	.80	1.20	2.40
	Vol. AU-1, No. 4, July–August 1953 (19 pages)	.70	1.05	2.10
	Vol. AU-1, No. 5, September–October 1953 (11 pages)	.50	.75	1.50
	Vol. AU-1, No. 6, November–December 1953 (27 pages)	.90	1.35	2.70
	Vol. AU-2, No. 1, January–February 1954 (38 pages)	1.20	1.80	3.60
	Vol. AU-2, No. 2, March–April 1954 (31 pages)	.95	1.40	2.85
	Vol. AU-2, No. 3, May–June 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 4, July–August 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 5, September–October 1954 (22 pages)	.95	1.40	2.85
	Vol. AU-2, No. 6, November–December 1954 (24 pages)	.80	1.20	2.40
	Vol. AU-3, No. 1, January–February 1955 (20 pages)	.60	.90	1.80
	Vol. AU-3, No. 2, March–April 1955 (51 pages)	.95	1.40	2.85
	Vol. AU-3, No. 3, May–June 1955 (85 pages)	.85	1.25	2.55
Broadcast Transmission Systems	PGBTS-1: March 1955 (102 pages)	2.50	3.75	7.50
Broadcast and Television Receivers	PGBTR-1: Round-Table Discussion on UHF TV Receiver Considerations, 1952 IRE National Convention (12 pages)	.50	.75	1.50
	PGBTR-3: June 1953 (67 pages)	1.40	2.10	4.20
	PGBTR-5: January 1954 (96 pages)	1.80	2.70	5.40
	PGBTR-6: April 1954 (119 pages)	2.35	3.50	7.00
	PGBTR-7: July 1954 (58 pages)	1.15	1.70	3.45
	PGBTR-8: October 1954 (20 pages)	.90	1.35	2.70
	Vol. BTR-1, No. 1, January 1955—Papers presented at the Radio Fall Meeting, 1954 (68 pages)	1.25	1.85	3.75
Circuit Theory	PGCT-1: IRE Western Convention, August 1952 (100 pages)	1.60	2.40	4.80
	PGCT-2: Papers presented at the Circuit Theory Sessions of the Western Electronic Show & Convention, San Francisco, Calif., August 19–21, 1953 (106 pages)	1.95	2.90	5.85
	Vol. CT-1, No. 1, March 1954 (80 pages)	1.30	1.95	3.90
	Vol. CT-1, No. 2, June 1954 (39 pages)	1.00	1.50	3.00
	Vol. CT-1, No. 3, September 1954 (73 pages)	1.00	1.50	3.00
	Vol. CT-1, No. 4, December 1954 (42 pages)	1.00	1.50	3.00

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(Continued)

TRANSACTIONS OF IRE PROFESSIONAL GROUPS

Sponsoring Group	Publications	Group Members	IRE Members	Non-* Members
	Vol. CT-2, No. 1, March 1955 (106 pages)	$2.70	$4.05	$8.10
Communications Systems	Vol. CS-2, No. 1, January 1954 (83 pages)	1.65	2.50	4.95
	Vol. CS-2, No. 3, November 1954: IRE Symposium on Global Communications, June 23–25, 1954, Washington, D. C. & IRE-AIEE Symposium on Military Communications, April 28, 1954, New York, N. Y. (181 pages)	3.00	4.50	9.00
	Vol. CS-3, No. 1, March 1955: Papers presented at the Symposium on Marine Communications & Navigation, October 13–15, 1954, Boston, Mass. (72 pages)	1.00	1.50	3.00
Component Parts	PGCP-1: March 1954 Issue (46 pages)	1.20	1.80	3.60
	PGCP-2: September 1954: Papers presented at the Component Parts Sessions at the 1954 Western Electronic Show & Convention, Los Angeles, Calif. (118 pages)	2.25	3.35	6.75
	PGCP-3: April 1955 (44 pages)	1.00	1.50	3.00
Electronic Computers	Vol. EC-2, No. 2, June 1953 (27 pages)	.90	1.35	2.70
	Vol. EC-2, No. 3, September 1953 (27 pages)	.75	1.10	2.25
	Vol. EC-2, No. 4, December 1953 (47 pages)	1.25	1.85	3.75
	Vol. EC-3, No. 1, March 1954 (39 pages)	1.10	1.65	3.30
	Vol. EC-3, No. 2, June 1954 (65 pages)	1.65	2.45	4.95
	Vol. EC-3, No. 3, September 1954 (54 pages)	1.80	2.70	5.40
	Vol. EC-3, No. 4, December 1954 (46 pages)	1.10	1.65	3.30
	Vol. EC-4, No. 1, March 1955 (48 pages)	1.10	1.65	3.30
	Vol. EC-4, No. 2, June 1955 (86 pages)	.90	1.35	2.70
Electron Devices	PGED-4: December 1953 (62 pages)	1.30	1.95	3.90
	Vol. ED-1, No. 1, February 1954 (72 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 2, April 1954 (75 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 3, August 1954 (77 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 4, December 1954—Papers presented at the 1954 Symposium on Fluctuation Phenomena in Microwave Sources, November 18–19, 1954, New York, N. Y. (280 pages)	3.20	4.80	9.60
	Vol. ED-2, No. 2, April 1955 (53 pages)	2.10	3.15	6.30
Engineering Management	PGEM-1: February 1954 (55 pages)	1.15	1.70	3.45
	PGEM-2: November 1954 (67 pages)	1.30	1.95	3.90
	PGEM-3: March 1955 (52 pages)	1.00	1.50	3.00
Industrial Electronics	PGIE-1: August 1953 (40 pages)	1.00	1.50	3.00
	PGIE-2: March 1955 (81 pages)	1.90	2.85	5.70
Information Theory	PGIT-2: A Bibliography of Information Theory (Communication Theory-Cybernetics)—(60 pages)	1.25	1.85	3.75
	PGIT-3: March 1954 (159 pages)	2.60	3.90	7.80
	PGIT-4: September 1954 (234 pages)	3.35	5.00	10.00
	Vol. IT-1, No. 1, March 1955 (76 pages)	2.40	3.60	7.20
Instrumentation	PGI-2: Data Handling Systems Symposium: IRE Western Electronic Show & Convention, Long Beach, Calif., August 27–29, 1952 (111 pages)	1.65	2.45	4.95
	PGI-3: April 1954 (55 pages)	1.05	1.55	3.15
Microwave Theory & Techniques	Vol. MTT-1, No. 2, November 1953 (44 pages)	.90	1.35	2.70
	Vol. MTT-2, No. 2, July 1954 (67 pages)	1.25	1.85	3.75
	Vol. MTT-2, No. 3, September 1954: Papers presented at the Joint IRE Professional Group—URSI meeting, Washington, D. C., May 5, 1954 (54 pages)	1.10	1.65	3.30
	Vol. MTT-3, No. 1, January 1955 (47 pages)	1.50	2.25	4.50
	Vol. MTT-3, No. 2, March 1955—Symposium on Microwave Strip Circuits, October 11–12, 1954, Tufts College, Medford, Mass. (182 pages)	2.70	4.05	8.10
	Vol. MTT-3, No. 3, April 1955 (44 pages)	1.40	2.10	4.20
Nuclear Science	Vol. NS-1, No. 1, September 1954 (42 pages)	.70	1.00	2.00
	Vol. NS-2, No. 1, June 1955 (15 pages)	.55	.85	1.65
Reliability & Quality Control	PGQC-2: March 1953 (51 pages)	1.30	1.95	3.90
	PGQC-3: February 1954 (39 pages)	1.15	1.70	3.45
	PGQC-4: December 1954 (56 pages)	1.20	1.80	3.60
	PGRQC-5: April 1955 (56 pages)	1.15	1.75	3.45

* Public Libraries, Colleges and Subscription Agencies may purchase at IRE Member rate.

(Continued)

Radio Pioneers—Attention!

This notice, from the Radio Pioneers Club, is for the information of members of the IRE who are qualified and wish to join the Radio Pioneers but who lack information concerning it. Originally formed in 1942 as the Twenty Year Club, the purposes of the organization are:

"To establish a membership organization of persons who by their long years of service in the field of Radio desire to become associated for the purposes of friendship and education. The Club shall be a central clearing house for the exchange of information and historical data about the radio industry and shall record in form to be determined facts and data about the history of the radio industry and its traditions for use by this and future generations. It is felt that this organization with the resultant exchange of information would make a valuable contribution to the public interest."

As outlined in the constitution, the purposes are broad enough to enable the Radio Pioneers to undertake almost any task which the club may desire for the furtherance of the radio industry.

1950 marked the establishment of a Radio Hall of Fame in which the memories of men and women whose contribution have placed them among the immortals will be perpetuated. The first of such awards went to Thomas Alvah Edison. Subsequent awards went to Marconi, Fessenden, Frank Conrad and Joseph Henry. All members are entitled to submit their choice which will be voted upon by the committee.

In 1950 the Radio Pioneers inaugurated the radio oral history project which has been undertaken by Columbia University's oral history project, under the direction of Allen Nevins. This vast project set up a source of material which will be available to all future historians of the industry, or to those writing books which dwell on some phases of radio. The plan consisted of obtaining recorded interviews of outstanding men and women who have pioneered in radio, engineers as well as those in other branches. It is felt that soon many of these will no longer be available for first-hand information. All members are encouraged to send in early anecdotes and "firsts" with which they were connected or concerned.

The requirements for membership are that a prospective member be in good standing, and have served the radio industry for 20 consecutive years or more at the time of making application for membership.

H. V. Kaltenborn, noted commentator and author, is the founder of the Pioneers; this came about when NBC gave him a dinner on April 4, 1942, to commemorate his 20th year in radio. The small original group has grown to more than 1000. Officers include Raymond F. Guy, formerly IRE President, President; John Patt (Station WJR Detroit) 1st Vice President; Lewis H. Avery (Avery-Knodel, Inc.) Vice-President; Victor Diehm (Station WAZL) Vice-President; and Merle Jones (Columbia Broadcasting System) Vice-President. Honorary Presidents

(Cont'd on page 1144)

TRANSACTIONS OF IRE PROFESSIONAL GROUPS

(Continued)

Sponsoring Group	Publications	Group Members	IRE Members	Non-* Members
Telemetry and Remote Control	PGRTRC-1: August 1954 (16 pages)	$ .85	$1.25	$2.55
	PGRTRC-2: November 1954 (24 pages)	.95	1.40	2.85
	Vol. TRC-1, No. 1, February 1955 (24 pages)	.95	1.40	2.85
Ultrasonics Engineering	PGUE-1: June 1954 (62 pages)	1.55	2.30	4.65
	PGUE-2: November 1954 (43 pages)	1.05	1.55	3.15
	PGUE-3: May 1955 (70 pages)	1.45	2.20	4.35
Vehicular Communications	PGVC-3: Theme—Spectrum Conservation, Washington, D. C., December 3–5, 1952 (140 pages)	3.00	4.50	9.00
	PGVC-4: Design, Planning & Operation of Mobile Communications Systems (June 1954) (98 pages)	2.40	3.60	7.20
	PGVC-5: June 1955 (76 pages)	1.50	2.25	4.50

* Public Libraries, Colleges and Subscription Agencies may purchase at IRE Member rate.

Radio Pioneers—Attention!

(*Cont'd from page 1143*)

are Brig. Gen. David Sarnoff and Dr. Lee De Forest.

Active local chapters exist in New York and Honolulu and plans are made to form others to promote friendship and social intercourse among the pioneer radio fraternities. New members have an opportunity to participate actively in the formation of their local chapters.

Radio Pioneers issues a small magazine and whenever feasible an annual membership roster including short biographies of each member.

Radio Pioneers holds its annual election, meeting and banquet each year at the time and place of the NARTB Convention and during the year a number of chapter meetings are held.

Membership dues in the Pioneers are $10 annually. The initiation cost is $15 each, $10 for dues and $5 for a Pioneers insignia pin.

The Radio Pioneers is the only association of its kind in that it is not confined to engineers, financial people, the legal fraternity, program people, or management personnel, but includes them all. The meetings provide an opportunity to create friendships and enjoy informal social affairs with fellow members representing all of the industry. The opportunity to meet and enjoy the companionship with pioneers in the non-engineering professions has been pleasant, stimulating and valuable.

Chapter meetings are highly informal and normally include excellent entertainment following dinner. Members of the Institute who are eligible to join are invited to send for an application blank to Raymond F. Guy, President, c/o National Broadcasting Company, Inc., 30 Rockefeller Plaza, New York 20, New York.

I.I.T. Will Sponsor Conference on Industrial Hydraulics

The use of electronic analog computers in the solution of hydraulic problems will be a feature of the eleventh annual National Conference on Industrial Hydraulics, October 27 and 28, at the La Salle Hotel, Chicago.

The conference will be sponsored by the Illinois Institute of Technology graduate school and Armour Research Foundation, in cooperation with engineering societies and nearly 100 industrial organizations.

Two Hundred Attend Philadelphia Student Night

The Philadelphia Section last spring organized a Student Night which included a dinner and meeting at the Engineers Club in Philadelphia. Nearly 100 students from the University of Delaware, Drexel Institute, Lafayette College, Lehigh University, University of Pennsylvania, Swarthmore, and Villanova were guests of the Section. After the dinner, attended by 200, a Section meeting was held. John Ryder, IRE President, spoke to the group on "Automatic Electronic Production." Also addressing the meeting was Theodore Hunter, Editor of the *Student Quarterly*. Student Awards were presented to seven students by S. C. Spielman, Chairman of Philadelphia Section.

Philadelphia Section Presents Student Awards to Members from Six Colleges

Winners of Student Awards at the Philadelphia Section meeting are (*standing, left to right*): L. A. Rubin, Pennsylvania; E. J. Taylor, Delaware; William Fryer, Lafayette; William Kilpatrick, Lafayette; C. F. Der, Drexel. *Seated:* T. S. Durand, Villanova; S. C. Speilman, Philadelphia Section Chairman; R. J. Fulmer, Lehigh.

Professional Group News

New York Chapter of PGANE Will Expand Winter Program

Now in its second full season, the New York Chapter of the Professional Group on Aeronautical and Navigational Electronics is scheduling an expanded program of six meetings. Functioning jointly for the New York, Long Island, and Northern New Jersey Sections, the chapter will cover topics in communications, control, navigation, and training aids.

The season will get under way on Thursday evening, October 20, with a session on "The Automatic Dead Reckoning Computer." Subsequent meetings and field trips will take place in November, January, February, April, and June.

Planning has been carried out by the Chapter Executive Committee, consisting of *Gordon P. McCouch*, Aircraft Radio Corp., Chairman; *Lester M. Glantz*, Telephonics Corp., Vice-Chairman; *William P. McNally*, W. L. Maxson Corp., Secretary; *Stamates I. Frangoulis*, Ford Instrument Co., Membership Committee Chairman; *Henry C. Nelson*, Polytechnic Research and Development Corp., Program Committee Chairman; and *Robert J. Bibbero*, Hillyer Instrument Co., Past Chairman. The Program Committee includes: *Thomas W. Winternitz*, Bell Telephone Labs.; *Robert A. Buckles*, Watson, Leavenworth, Kelton and Taggart; *Charles Cambridge*, N. Y. USAF Development Field Office; *Richard Meyers*, Federal Telecommunications Labs.; and *John Litsios*, W. L. Maxson Corp.

Four New Chapters Announced

On May 4, the IRE Board of Directors approved the establishment of a new Section in Ontario, Canada, to be known as the Bay of Quinte Section.

At the same meeting, the Amarillo-Lubbock Subsection of the Dallas-Fort Worth Section was made a full Section, now called the Lubbock Section.

The Executive Committee, at its meeting on June 7, approved the following chapters: Central Florida Section, PG on Telemetry and Remote Control; Philadelphia Section, PG on Medical Electronics; Boston Section, PG on Automatic Control; Rome-Utica Section, PG on Communications Systems.

Obituary

James F. Pierce (A'30–VA'39) died recently. Mr. Pierce, a patent attorney, received the bachelor's degree in civil engineering from the University of Michigan and the bachelor of laws degree from George Washington University. He was a World War I veteran and a member of the American Legion. He also held membership in the American Bar Association and the American Patent Law Association. He had been admitted to practice before the Supreme Court, and was a partner in the patent law firm of Pierce, Scheffler and Parker, Washington, D. C.

Mr. Pierce was an IRE Representative at the University of Pittsburgh in 1949 and 1950.

Technical Committee Notes

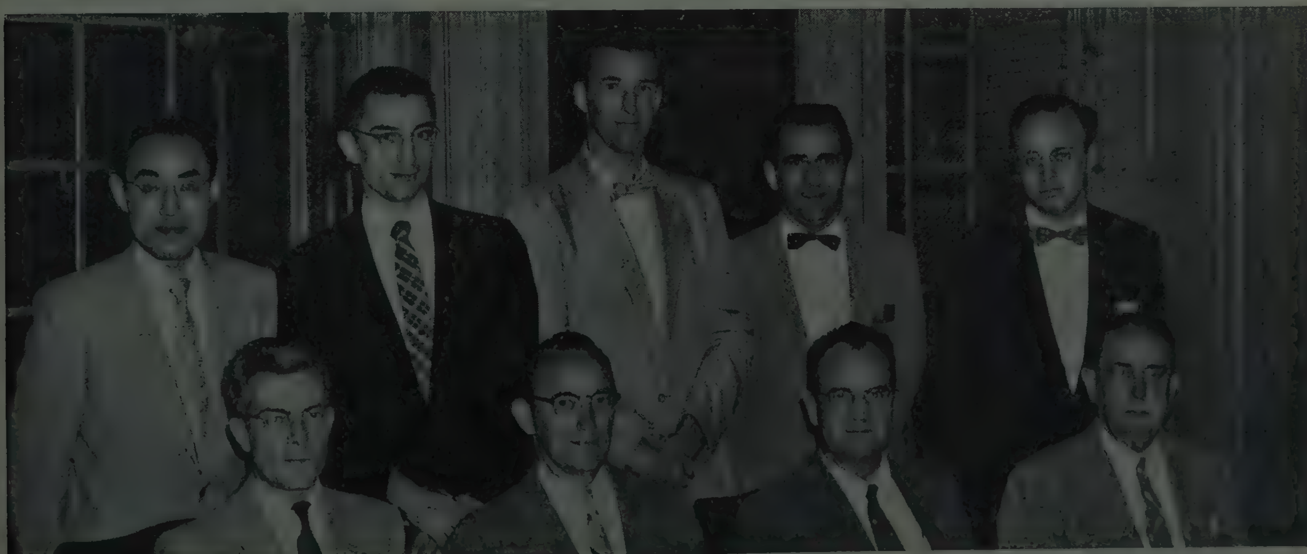
The **Antennas and Waveguides** Committee met at IRE Headquarters on June 8 with Henry Jasik presiding. Mr. Jasik reported on the Standards Committee review of the Proposed Definitions of Waveguide Components. The work of the Waveguide and Waveguide Component Measurements Subcommittee (2.4) on Methods of Waveguide Measurement was reviewed. The sections on delay time and power handling capacity had no further change. The section on Q was discussed at length and amended. This Proposed Standard will be discussed further at the next meeting.

W. R. Bennett presided at a joint meeting of the **Circuits** Committee and the Subcommittee on Linear Active Circuits Including Network with Feedback Servomechanism (4.7) on June 23 at IRE Headquarters. It was announced that S. J. Mason had resigned from the Circuits Committee because of the pressure of other work, and that W. A. Lynch would take his place as Vice-Chairman. The committee discussed, amended, and approved the feedback definitions for submission to the Standards Committee.

The **Facsimile** Committee met at the Times Building on June 17 with K. R. McConnell presiding. Mr. Lankes reported on the progress of the IRE Facsimile Standards Chart from a series of prepared notes. The committee decided to write down a short definition and use for each pattern of the chart. The committee discussed the facsimile definitions as sent to the Standards Committee.

Mr. J. E. Eiselein presided at a meeting of the **Industrial Electronics** Committee at IRE Headquarters on June 15. The chairman announced that the definitions on induction and dielectric heating were approved by the Standards Committee with some short added notes on the two words, decalesence and recalesence. The secretary was instructed to send a letter to R. R. Batcher, Chairman of the Professional Group on Production Techniques. The letter was to stipulate that those in the Industrial Electronics Committee had assumed work in the production techniques and automation fields and that they would appreciate it if the Production Techniques group worked with them. The letter also indicated a desire to have any interested members of this professional group as members of the Industrial Electronics Committee and a desire to have them recognize the group as their channel of expression. Mr. Cottle reported on the RETMA work on definitions. Definitions for

Administrative Committee of PG on Microwave Theory and Techniques Meets at IRE



At a recent meeting of the Administrative Committee of the PG on Microwave Theory and Techniques A. C. Beck was elected Chairman of the Group. Pictured at IRE Headquarters are (*seated left to right*): H. F. Engelmann, Vice-Chairman; A. C. Beck, Chairman; W. W. Mumford, former Chairman; G. C. Southworth, Chairman of Awards Committee. Standing (*left to right*) are: K. Tomiyasu, Secretary-Treasurer; R. F. Schwartz, who organized the Philadelphia Chapter; S. D. Schreyer, Secretary of the Baltimore Chapter; T. S. Saad, Editor; and Henry Jasik, of the Papers Review Committee. The meeting was held early this summer.

automatic machinery, aspect ratio and automatic dip soldering, prepared by Mr. Eiselein, were discussed and some modifications suggested. For those terms containing "automatic" it was decided to accept the dictionary definition for automatic and define the rest of the terms separately. It was decided to appoint a small task group to work on definitions of other terms and present them for consideration at the next meeting.

The **Information Theory and Modulation Systems** Committee met at IRE Headquarters on June 1 with J. G. Kreer presiding. The committee discussed and amended the Information Theory Definitions now under consideration.

H. R. Mimno presided at a meeting of the **Navigation Aids** Committee at IRE Headquarters on June 17. The committee continued its discussion of the proposed standard on "VHF Omni-Directional Radio Range (VOR)." Minor revisions and minor editorial changes were noted. The committee noted, but did not settle, certain policy questions regarding the possible subdivision of the system into major components for measurement standardization purposes.

The **Standards Committee** met at IRE Headquarters on June 7 with Chairman E. Weber presiding. The following proposed standards were discussed, amended, and approved and will appear in THE PROCEEDINGS shortly: "Standards on Industrial Electronics: Definitions of Industrial Electronic Terms, 1955;" "Standards on Antennas and Waveguides: Definitions of Waveguide Components, 1955;" "Standards on Pulses: Methods of Measurement of Pulse Quantities, 1955;" "Standards on Radio Receivers: Method of Testing Receivers Employing Ferrite Core Loop Antennas, 1955;" and "Standards on Graphical and Letter Symbols for Feedback Control Systems, 1955." The proposed standard on "Terminology for Feedback Control Systems" was discussed and will be completed at the next meeting of the Standards Committee. It was unanimously approved on motion by Mr. Shea and seconded by Mr. Baldwin that the Standards Committee recommend to the Executive Committee that the title of the Semiconductor Devices Committee be changed to the Solid State Devices Committee and their scope be changed accordingly. Dr. Weber announced that the following appointments had been approved at the last meeting of the Executive Committee: upon resignation of Mr. Dodds, P. A. Redhead was appointed Chairman of the Electron Tubes Committee; L. E. Coffey, Telecommunication Division, Department of Transport, Ottawa, Canada was appointed as a non-IRE member to the Radio Frequency Interference Committee. P. A. Fleming, British Radio Valve Manufacturers, London, England, was appointed as a non-IRE member to the Electron Tubes Committee.

Books

Analog Methods in Computation and Simulation by Walter W. Soroka

Published (1954) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 380 pages+9 page index+xii pages. Illus. 9¼×6¼. $7.50.

This book describes various electrical and mechanical components which (ideally) obey fundamental mathematical laws, and proceeds to demonstrate how computers and simulators may be constructed from these basic building blocks. The chapters are: "Mechanical Computing Elements;" "Electromechanical, Electrical, and Electronic Computing Elements;" "Machines for Simultaneous Linear Algebraic Equations;" "Analog Solution of Nonlinear Algebraic Equations;" "The Mechanical Differential Analyzer;" "Electronic Analog Computers (Electronic Differential Analyzers);" "Dynamical Analogies;" "Equivalent Circuits for Ordinary and Partial Differential Equations in Finite Differences;" and "Membrane and Conducting-sheet Analogies."

The analog computer has both a long history and a current vitality. The basic precepts may be quickly enumerated, but a feeling for the subject is developed only through acquaintance with a wide spectrum of examples, many in the "ingenious device" category. Prof. Soroka has provided these examples in profusion, yet without making the book a mere catalog of miscellany. He has done a good job of collecting and organizing material largely available heretofore only in isolated spots. The book is descriptive, but the author does not hesitate to "put the numbers in."

The problem of scaling receives adequate attention throughout the book, and the technical limitations of many of the computers described are indicated. A particularly valuable addition would have been the expansion of the brief chapter on electronic analog computers to include checking procedures and error analysis techniques.

The author's preface indicates that the book is considered a textbook; however, no exercises for the student are included. Properly employed, it could form the basis for a course at perhaps a high undergraduate level, although appreciation of the fine points in certain chapters requires somewhat more advanced training. Laboratory work would be essential, and might well be modeled upon examples in the book.

The practicing engineer will find this a useful basic reference, but one which does not attempt to solve his practical problems of detail design to meet space and weight limitations or to ensure reliable operation under adverse environmental conditions. In this connection, the reader will be grateful that the book contains a wealth of basic references to the literature.

LOUIS B. WADEL
Chance Vought Aircraft, Inc.
Dallas, Texas

Servomechanisms and Regulating System Design: Volume II by Harold Chestnut and Robert Mayer

Published (1955) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 1, N. Y. 368 pages+8 page index+xii pages. Illustrated. 9¼×6. $8.50.

An engineer who has had the experience of reducing a control system to practice soon recognizes the rather substantial gap between theory and practical design. Theory considers the broad and general fundamentals that describe the performance of the system. On the other hand, practical design requires the translation of the control task into technical specifications and the specific means of implementation. Included in the latter are error-sensing circuits, actuators, transducers, amplifiers and other components. The designer must also recognize and minimize the effects of imperfections such as drift, backlash, noise, saturation, nonlinearities, and overheating which are present in the physical components he is compelled to use. While there are many books which deal with the theory of feedback control systems (such as volume one of this same work), there are very few which deal with practical design. For this reason, this book is a most valuable and welcome addition to the library of the control engineer.

The first part deals with the measurement techniques for obtaining transfer functions of components The only criticism offered here is that this reviewer's experience has been that except for simple cases the accurate deduction of transfer functions from test data is not quite as direct and simple as the authors seem to imply. Following this, there is material on the setting of dynamic specifications of the system as influenced by the systematic inputs the system is required to follow and the noise it is required to reject. Selection of motors and actnators, network design, amplifier design and the handling of ac carrier servomechanisms round out the material on design. The authors restrict themselves largely to electrical and electronic systems, omitting other forms of implementation. It concludes with a substantial amount of material on nonlinearities such as saturation and backlash, on-off systems and the use of non-linear components for compensation.

Generally speaking, the book is well written and is replete with information of practical value to the engineer. The presentation, while analytical, is not too complex thus making the book easily readable by the practicing engineer and student. A comprehensive bibliography is available as an appendix so that the reader may broaden his coverage should he wish to do so. The lack of problems is one factor which lessens its value as a student textbook. On the other hand, it is recognized that meaningful problems in practical design are very difficult to devise and the lack of problems is quite understandable. This book is a very worthwhile addition to the list of texts dealing with feedback control and it is highly recommended to those engineers and students who have an interest in the practical design of servomechanisms.

JOHN R. RAGAZZINI
Columbia University
New York, N. Y.

The Amplification and Distribution of Sound by A. E. Greenlees

Published (1954) by Chapman and Hall Ltd., 37 Essex St. W.C. 2, London, England. 295 pages+5 page index+x pages. 114 Figs. 8¾×5⅝. 35s.

The author of this third revised edition faithfully pursues the purpose stated in the preface—"to present a general survey of the principles of sound amplification and distribution, showing the practical considerations involved, together with sufficient technical detail to enable the reader to appreciate the fundamental principles." The book is a survey in that a wide variety of subjects is covered lightly, though not superficially. By reading this book a beginner in sound amplification covers the field rapidly without getting deeply involved in any one branch of the subject. The emphasis throughout is on practical consideration and, although many technical details are omitted, it is to the author's credit that technical correctness has not been sacrificed.

Chapters are included on amplifier components, typical amplifier circuits, amplifier performance factors; audio sources such as radio receivers, microphones, and records of various types; loudspeakers, installation acoustics, system layouts; operation and maintenance of equipment; and the preparation of sound equipment specifications. No attempt is made to assign specific values to circuit components or to describe the mechanical design of electroacoustic components in terms of actual dimensions. The numerous block diagrams, charts, and schematics are intended to be illustrative rather than specific.

Little, if any, mathematical background or prerequisite reading is required for the reader of this book. The reader would do well, however, to have as alternative background a modicum of experience with sound systems, and additional experience of this type is recommended as an accompaniment to the study program.

Although sound amplification systems cannot, as the author states, "improve the acoustics of the building," their proper design, installation, and use can go far to overcome an adverse acoustical environment. The book will be very useful to sound system dealers, service technicians, and operators, as well as to architects and others in need of immediate direct information on sound amplification systems.

DANIEL W. MARTIN
Baldwin Piano Company
Cincinnati, Ohio

An Outline of Atomic Physics by O. H. Blackwood, T. H. Osgood, and A. E. Ruark

Published (1955) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 474 pages+12 page index+x pages+9 Appendixes. Illus. 9¼×6. $7.50.

This well-rounded exposition is the third edition of a text originally designed to provide college students with a thorough-going knowledge and understanding of the structure and behavior of atoms, molecules, and radiation. While the intended reader is expected to have completed a year's course in college physics, the book is written primarily for students aspiring to professions other than physics.

The contents of the book may be roughly divided into two classes. First nine chapters contain descriptions of the atomic nature of matter and of electricity, the forms and properties of radiant energy, atomic and molecular spectra, the duality of waves and particles, and a brief treatment of solids. The remaining six chapters may be roughly classed as nuclear physics, and include discussions of radioactivity, the elementary particles, nuclear transmutations, cosmic rays, and the theory of relativity.

The authors have avoided the use of calculus and higher mathematics throughout the text; nevertheless they make a determined (and generally highly successful) attempt to give the full story of the physics involved. They give to the reader not only the present-day physical theories but also the ideas which led to their formation. Where our knowledge is vague or contradictory this is pointed out, and the dilemmas of physics are discussed as freely as its successes.

Altogether, the radio engineer who is interested in getting an accurate and complete picture of the physics which interrelates the phenomena underlying his profession, and who wishes, as the authors state, "to go to the frontiers of physics and see for himself what manner of things must be done to take the next step forward" will find this book rewarding. The subject matter is not easy, and most readers will find that a considerable degree of concentration is required. The authors' presentation is, however, remarkable for its clarity, and those who expend the effort to absorb and retain this material will find unfolding a fascinating story of the triumphs of modern thought in unravelling some of the fundamental mysteries of nature.

JOHN W. COLTMAN
Westinghouse Research Laboratories
East Pittsburgh, Pennsylvania

Servomechanism Practice by William R. Ahrendt

Published (1954) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 341 pages+7 page index+vii pages. 283 figures. 9¼×6¼. $7.00.

This book is ideal for a manufacturer of instrument servomechanisms. The solutions which it gives to problems in fabrication, construction, component selection, circuitry, vibration, temperature variations, tolerances, and nonlinearities, are of considerable value to maintenance engineers and to manufacturers of all types of feedback control systems.

The writing is clear, simple, nonmathematical, easy to read, and accurate. It can be studied and mastered by a practicing engineer or an undergraduate student without the help of a teacher. Block diagrams are used effectively in the presentation by often appearing with the detailed circuit diagram.

There are a number of good examples of complete electric and hydraulic instrument servomechanisms with specifications, component characteristics, system analyses, and actual system performances. The book is excellent in its treatment of manufacturing techniques, nonlinearities in small components, and the observed effect of these non-linearities on system performance. The selection of a prime mover and gear ratios is covered in detail.

In its limited scope, this work fulfills an important need, duplicating no other book. It has very little, however, on control theory, high power industrial control systems, analogs or function computers. It does not discuss process control servos, guidance of large objects, digitally or periodically controlled systems, on-off or relay systems, nonlinear predictor systems, tension, velocity, or fabrication controls. It has nothing on statistical disturbances, or the formulation of specifications. It does not cover graphical, electrolytic, or analog aids for the design steps.

This book unfortunately does not have a wide enough scope to serve as a laboratory text for engineers in training, but it is an excellent guide for the final design and production engineers who must bridge the gap between preliminary design and delivery.

OTTO J. M. SMITH
Instituto Tecnologico de Aeronautica,
Sao Jose dos Campos, Est. Sao Paulo, Brasil

RECENT BOOKS

ASTM Standards on Electrical Insulating Materials. Compiled by ASTM Committee D-9 on Electrical Insulating Materials. American Society for Testing Materials, 1916 Race St., Philadelphia 3, Pa. $5.50.

Banner, E. H. W., *Electronic Measuring Instruments.* Chapman and Hall Ltd., 37 Essex St., W.C. 2, London, England. 45 *S.*

Duschinsky, W. J., *TV Stations: A Guide for Architects, Engineers, and Management.* Reinhold Publishing Corp., 430 Park Ave., N. Y. 30, N. Y. $12.00.

Higdon, Archie, and Stiles, William B., *Engineering Mechanics.* Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. $7.95.

Kiver, Milton S., *Introduction to UHF Circuits and Components.* D. Van Nostrand Co., Inc., 250 Fourth Ave., New York, N. Y. $7.50.

Proceedings of the Third Meeting of the Joint Commission on Radiometeorology. U.R.S.I. 42 Rue des Minimes, Brussels, Belgium. $5.00.

Professional Groups

Aeronautical & Navigational Electronics—*Chairman*, Edgar A. Post, Navigational Aides, United Air Lines, Operations Base, Stapleton Field, Denver 7, Colo.

Antennas & Propagation—*Chairman*, Delmer C. Ports, Jansky & Bailey, 1339 Wisconsin Ave., N.W., Washington 7, D. C.

Audio—*Chairman*, W. E. Kock, Bell Tel. Labs., Murray Hill, N. J.

Automatic Control—*Chairman*, Robert B. Wilcox, Raytheon Manufacturing Co., 148 California St., Newton 58, Mass.

Broadcast & Television Receivers—*Chairman*, W. P. Boothroyd, Philco Corp., Philadelphia 34, Pa.

Broadcast Transmission Systems—*Chairman*, O. W. B. Reed, Jr., Jansky & Bailey, 1735 DeSales St., N.W., Washington, D. C.

Circuit Theory—*Chairman*, J. Carlin, Microwave Res. Inst., Polytechnic Inst. of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.

Communications Systems—*Chairman*, A. C. Peterson, Jr., Bell Labs., 463 West St., New York 14, N. Y.

Component Parts—*Chairman*, Floyd A. Paul, Reliability Bendix Development Lab., 116 W. Olive Avenue, Burbank, Calif.

Electron Devices—*Chairman*, J. S. Saby, Electronics Laboratory, G.E. Co., Syracuse, N. Y.

Electronic Computers—*Chairman*, J. H. Felker, Bell Labs., Whippany, N. J.

Engineering Management—*Chairman*, C. J. Breitwieser, Lear, Inc., 3171 S. Bundy Drive, Los Angeles 34, Calif.

Industrial Electronics—*Chairman*, George P. Bosomworth, Engrg. Lab., Firestone Tire & Rubber Co., Akron 17, Ohio

Information Theory—*Chairman*, Louis A. DeRosa, Federal Telecommunications Lab., Inc., 500 Washington Avenue, Nutley, N. J.

Instrumentation—*Chairman*, F. G. Marble, Boonton Radio Corp., Intervale Rd., Boonton, N. J.

Medical Electronics—*Chairman*, Dr. Julia F. Herrick, Inst. of Experimental Medicine, Mayo Found, Rochester, Minn.

Microwave Theory and Techniques—*Chairman*, A. C. Beck, Bell Labs., 463 West St., New York 14, N. Y.

Nuclear Science—*Chairman*, M. A. Schultz, Westinghouse Automatic Power Division, Bettis Field, Pittsburgh 30, Pa.

Reliability and Quality Control—*Chairman*, Victor Wouk, Beta Electric Corp., 333 E. 103rd St., New York 29, N. Y.

Production Techniques—*Chairman*, R. R. Batcher, 240-02—42nd Ave., Douglaston, L. I., N. Y.

Telemetry and Remote Control—*Chairman*, C. H. Hoeppner, Stavid Engineering, Plainfield, N. J.

Ultrasonics Engineering—*Chairman*, M. D. Fagen, Bells Labs., Whippany, N. J.

Vehicular Communications—Newton Monk, Bell Telephone Labs., 463 West St., New York, N. Y.

Sections*

Akron (4)—H. L. Flowers, 2029 19 St., Cuyahoga Falls, Ohio; H. F. Lanier, 49 W. Lowell Ave., Akron, Ohio.

Alberta (8)—Officers to be elected.

Albuquerque-Los Alamos (7)—T. G. Banks, Jr., 1124 Monroe St., S.E., Albuquerque, N. M.; G. A. Fowler, 3333 49 Loop, Sandia Base, Albuquerque, N. M.

Atlanta (3)—D. L. Finn, School of Electrical Engineering, George Institute of Technology, Atlanta, Ga.; P. C. Toole, 605 Morningside Dr., Marietta, Ga.

Baltimore (3)—C. F. Miller, Johns Hopkins University, 307 Ames Hall, Baltimore 18, Md.; H. R. Hyder, III, Route 2, Owings Mills, Md.

Bay of Quinte (8)—J. C. R. Punchard, Elec. Division, Northern Elec. Company, Ltd., Sydney St., Belleville, Ont., Canada; M. J. Waller, R.R. 1, Foxboro, Ont., Canada.

Beaumont-Port Arthur (6)—W. W. Eckles, Jr., Sun Oil Company, Prod. Lab., 1096 Calder Ave., Beaumont, Tex.; E. D. Coburn, Box 793, Nederland, Tex.

Binghamton (4)—O. T. Ling, 100 Henry St., Binghamton, N. Y.; Arthur Hamburgen, 102 S. Nanticoke Ave., Endicott, N. Y.

Boston (1)—T. P. Cheatham, Jr., Hosmer St., Marlborough, Mass.; R. A. Waters, 4 Gordon St., Waltham 54, Mass.

Buenos Aires—J. M. Rubio, Ayachucho 1147, Buenos Aires, Argentina; J. L. Blon, Transradio Internacional, San Martin 379, Buenos Aires, Argentina.

Buffalo-Niagara (1)—D. P. Welch, 859 Highland Ave., Buffalo 23, N. Y.; W. S. Holmes, 1961 Ellicot Rd., West Falls, N. Y.

Cedar Rapids (5)—Ernest Pappenfus, 1101 30 St. Dr., S.E., Cedar Rapids, Iowa; E. L. Martin, 1119 23 St., S.E., Cedar Rapids, Iowa.

Central Florida (3)—Hans Scharla-Nielsen, Radiation Inc., P.O. Drawer "Q," Melbourne, Fla.; G. F. Anderson, Radiation Inc., P.O. Box "Q," Melbourne, Fla.

Chicago (5)—J. S. Brown, 9829 S. Hoyne Ave., Chicago 43, Ill.; D. G. Haines, 17 W. 121 Oak Lane, Bensenville, Ill.

Cincinnati (4)—D. W. Martin, The Baldwin Company, 1801 Gilbert, Cincinnati 2, Ohio; F. L. Wedig, Jr., 3819 Davenant Ave., Cincinnati 13, Ohio.

Cleveland (4)—R. H. DeLany, 5000 Euclid Ave., Cleveland 3, Ohio; J. F. Keithley, 22775 Douglas Rd., Shaker Heights 22, Ohio.

Columbus (4)—R. W. Masters, 1633 Essex Rd., Columbus 21, Ohio; R. L. Cosgriff, 2200 Homestead, Columbus, Ohio.

Connecticut Valley (1)—P. F. Ordung, Dunham Laboratory, Yale University, New Haven, Conn.; H. M. Lucal, Box U-37, University of Connecticut, Storrs, Conn.

Dallas-Fort Worth (6)—M. W. Bullock, 6805 Northwood Rd., Dallas 25, Tex.; C. F. Seay, Jr., Collins Radio Company, 1930 Hi-Line Dr., Dallas, Tex.

Dayton (5)—M. A. McLennan, 304 Schenck Ave., Dayton 9, Ohio; N. A. Nelson, 310 Lewiston Rd., Dayton 9, Ohio.

Denver (6)—J. W. Herbstreit, 2000 E. Ninth Ave., Boulder, Colo.; R. S. Kirby, 455 Hawthorne Ave., Boulder, Colo.

Des Moines-Ames (5)—A. A. Read, 511 Northwestern Ave., Ames, Iowa; W. L. Hughes, E.E. Department, Iowa State College, Ames, Iowa.

Detroit (4)—N. D. Saigeon, 1544 Grant, Lincoln Park 25, Mich.; A. L. Coates, 1022 E. Sixth St., Royal Oak, Mich.

Elmira-Corning (1)—J. L. Sheldon, 179 Dodge Ave., Corning, N. Y.; J. P. Hocker, Corning Glass Works, Corning, N. Y.

El Paso (6)—J. C. Nook, 1126 Cimarron St., El Paso, Tex.; J. H. Maury, 3519 Ft. Blvd., El Paso, Tex.

Emporium (4)—E. H. Boden, R.D. 1, Emporium, Pa.; H. S. Hench, Jr., R.D. 2, Emporium, Pa.

Evansville-Owensboro (5)—A. P. Haase, 2230 St. James Ct., Owensboro, Ky.; D. D. Mickey, Jr., Engineering Department, General Electric Company, Owensboro, Ky.

Fort Wayne (5)—C. L. Hardwick, 2905 Chestnut St., Fort Wayne 4, Ind.; Paul Rudnick, Farnsworth Electronics Company, Fort Wayne 1, Ind.

Hamilton (8)—G. F. Beaumont, 6 Tallman Ave., Burlington, Ont., Canada; C. N. Chapman, 40 Dundas St., Waterdown, Ont., Canada.

Hawaii (7)—H. E. Turner, 44-271 Mikiola Dr., Kaneohe, Hawaii; G. H. Hunter, Box 265, Lanikai, Oahu, T.H.

Houston (6)—L. W. Erath, 2831 Post Oak Rd., Houston, Tex.; J. M. Bricaud, Schlumberger Well Surveying Corporation, Box 2175, Houston 1, Tex.

Huntsville (3)—D. E. French, 1403 E. Clinton St., Huntsville, Ala.; T. L. Greenwood, 1709 La Grande St., Huntsville, Ala.

Indianapolis (5)—A. J. Schultz, 908 E. Michigan St., Indianapolis, Ind.; H. L. Wisner, 5418 Rosslyn Ave., Indianapolis 20, Ind.

Inyokern (7)—G. D. Warr, 213-A Wasp Rd., China Lake, Calif.; B. B. Jackson, 54-B Rowe St., China Lake, Calif.

Israel—Franz Ollendorf, Box 910, Hebrew Institute of Technology, Haifa, Israel; J. H. Halberstein, P.O. Box 1, Kiriath Mozkin, Israel.

Ithaca—Benjamin Nichols, School of Electrical Engineering, Cornell University, Ithaca, N. Y.; Howard Heydt, General Electric Co., Advanced Electronics Center, Cornell University Airport, Ithaca, N. Y.

* Numerals in parentheses following Section designate Region number. First name designate, Chairman, second name, Secretary.

(Cont'd on next page)

(Sections cont'd)

Kansas City (6)—K. V. Newton, Bendix Aviation Corporation, Box 1159, Kansas City 41, Mo.; Mrs. G. L. Curtis, Radio Industries, Inc., 1307 Central Ave., Kansas City 2, Kan.

Little Rock (6)—J. E. Wylie, 2701 N. Pierce, Little Rock, Ark.; J. C. Spilman, 34 Lakeshore Dr., Route 9, Little Rock, Ark.

London (8)—C. F. MacDonald, 328 St. James St., London, Ont., Canada; J. D. B. Moore, 27 McClary Ave., London, Ont., Canada.

Long Island (2)—P. G. Hansel, Addison Lane, Greenvale, L. I., N. Y.; J. Neidert, 9 Surrey Rd., New Hyde Park, L. I., N. Y.

Los Angeles (7)—B. S. Angwin, 3300 Colby Ave., Los Angeles 34, Calif.; C. E. Rutherford, 209 S. Oakhurst Dr., Beverly Hills, Calif.

Louisville (5)—R. W. Mills, 1017 Eastern Pkwy., Louisville 4, Ky.; L. A. Miller, 314 Republic Building, Louisville 2, Ky.

Lubbock (6)—H. A. Spuhler, E. E. Dept., Texas Tech. College, Lubbock, Tex.; J. W. Dean, 1903 49 St., Lubbock, Tex.

Miami (3)—C. S. Clemans, Station WSWN, Belle Glade, Fla.; H. F. Bernard, 1641 S.W. 82 Pl., Miami, Fla.

Milwaukee (5)—W. E. Watts, 2224 N. 70 St., Wauwatosa 13, Wis.; W. A. Van Zeeland, 4510 N. 45 St., Milwaukee 16, Wis.

Montreal (8)—Sydney Bonneville, Room 1427, 1050 Beaver Hall Hill, Montreal, P. Q., Canada; R. E. Penton, 2090 Claremont Ave., Montreal, P. Q., Canada.

New Orleans (6)—J. A. Cronvich, Department of Electrical Engineering, Tulane Univ., New Orleans 18, La.; N. R. Landry, 620 Carol Dr., New Orleans 21, La.

New York (2)—A. B. Giordano, 85-99 Livingston St., Brooklyn, N. Y.; H. S. Renne, Radio-Electronic Engr., 366 Madison Ave., New York 17, N. Y.

North Carolina-Virginia (3)—J. C. Mace, 1616 Jefferson Park Ave., Charlottesville, Va.; A. L. Comstock, 1404 Hampton Dr., Newport News, Va.

Northern New Jersey (2)—W. R. Thurston, 923 Warren Pkwy., Teaneck, N. J.; R. J. Kircher, 145 Maple St., Summit, N. J.

Northwest Florida (3)—B. H. Overton, Box 115, Shalimar, Fla.; G. C. Fleming, 579 E. Gardner Dr., Fort Walton Beach, Fla.

Olkahoma City (6)—A. P. Challenner, University of Oklahoma, Norman, Okla.; Frank Herrmann, 1913 N.W. 21 St., Oklahoma City, Okla.

Omaha-Lincoln (5)—M. L. McGowan, 5544 Mason St., Omaha 6, Neb.; C. W. Rook, Department of Electrical Engineering, University of Nebraska, Lincoln 8, Neb.

Ottawa (8)—George Glinski, 14 Dunvegan Rd., Ottawa, Ont., Canada; C. F. Pattenson, 3 Braemar, Ottawa 2, Ont., Canada.

Philadelphia (3)—C. R. Kraus, Bell Telephone Company of Pennsylvania, 1835 Arch St., Philadelphia 3, Pa.; Nels Johnson, Philco Corporation, 4700 Wissahickon Ave., Philadelphia 44, Pa.

Phoenix (7)—W. R. Saxon, 641 E. Missouri, Phoenix, Ariz.; G. L. McClanathan, 509 E. San Juan Cove, Phoenix, Ariz.

Pittsburgh (4)—J. N. Grace, 112 Heather Dr., Pittsburgh 34, Pa.; J. B. Woodford, Jr., Box 369, Carnegie Tech. P.O., Pittsburgh 13, Pa.

Portland (7)—J. M. Roberts, 4325 N.E. 77, Portland 13, Ore.; D. C. Strain, 7325 S.W. 35 Ave., Portland 19, Ore.

Princeton (2)—G. C. Sziklai, Box 3, Princeton, N. J.; L. L. Burns, Jr., R.C.A. Laboratories, Princeton, N. J.

Rochester (1)—Allan Holstrom, 551 Spencer Rd., Rochester 9, N. Y.; W. F. Bellor, 186 Dorsey Rd., Rochester 16, N. Y.

Rome-Utica (1)—Harry Davis, 716 Cherry St., Rome, N. Y.; M. V. Ratynski, 205 W. Cedar St., Rome, N. Y.

Sacramento (7)—R. C. Bennett, 3401 Chenu Ave., Sacramento, Calif.; R. A. Poucher, Jr., 3021 Mountain View Ave., Sacramento 21, Calif.

St. Louis (6)—F. A. Fillmore, 5758 Itaska St., St. Louis 9, Mo.; Christopher Efthim, 1016 Louisville Ave., St. Louis 10, Mo.

Salt Lake City (7)—M. E. Van Valkenburg, Department of Electrical Engineering, University of Utah, Salt Lake City 1, Utah; A. L. Gunderson, 3906 Parkview Dr., Salt Lake City, Utah.

San Antonio (6)—C. M. Crain, Engineering Building 149, University of Texas, Austin 12, Tex.; W. H. Hartwig, Dept. E. E., University of Texas, Austin 12, Tex.

San Diego (7)—F. X. Byrnes, 1759 Beryl St., San Diego 9, Calif.; R. T. Silberman, 4274 Middlesex Dr., San Diego, Calif.

San Francisco (7)—B. M. Oliver, 395 Page Mill Rd., Palo Alto, Calif.; Wilson Pritchett, Div. of Electrical Engineering, University of California, Berkeley 4, Calif.

Schenectady (1)—C. C. Allen, 2064 Baker Ave., Schenectady 9, N. Y.; A. E. Rankin, 833 Whitney Dr., Schenectady, N. Y.

Seattle (7)—W. C. Galloway, 5215 Pritchard St., Seattle 6, Wash.; L. O. Nelson, 10303 13 Ave., N.W., Seattle 77, Wash.

Syracuse (1)—A. D. Arsem, G.E. Co., Electronics Park, Syracuse, N. Y.; G. M. Glasford, E. E. Dept., Syracuse Univ., Syracuse 10, N. Y.

Toledo (4)—L. R. Klopfenstein, Portage, Ohio; D. F. Cameron, 1619 Milburn Ave., Toledo 6, Ohio.

Toronto (8)—A. P. H. Barclay, 2 Pine Ridge Dr., Toronto 13, Ont., Canada; H. W. Jackson, 352 Laird Dr., Toronto 17, Ont., Canada.

Tulsa (6)—C. F. Hadley, 1356 E. 45 Pl., Tulsa 15, Okla.; L. H. Hooker, 4064 E. 22 Pl., Tulsa 5, Okla.

Twin Cities (5)—N. B. Coil, 1664 Thomas Ave., St. Paul 4, Minn.; A. W. Sear, 5801 York Ave., S., Minneapolis 10, Minn.

Vancouver (8)—J. E. Breeze, 5591 Toronto Rd., Vancouver 8, B. C., Canada; R. A. Marsh, 3873 W. 23 Ave., Vancouver, B. C., Canada.

Washington (3)—H. I. Metz, U. S. Government Department of Commerce, CAA, Room 2076, T-4 Building, Washington 25, D. C.; A. H. Schooley, 3940 First St., S.W., Washington 24, D. C.

Williamsport (4)—J. E. Snook, 1629 Warren Ave., Williamsport, Pa.; F. T. Henry, 1345 Pennsylvania Ave., Williamsport, Pa.

Winnipeg (8)—R. M. Simister, 179 Renfrew St., Winnipeg, Man., Canada; G. R. Wallace, 400 Smithfield Ave., Winnipeg, Man., Canada.

Subsections

Amarillo-Lubbock (6)—R. B. Spear, 510 E. Hill St., Brownfield, Tex.; J. B. Joiner, 2621 30 St., Lubbock, Tex.

Berkshire (1)—Gilbert Devey, Sprague Elec. Co., Marshall St., Building 1, North Adams, Mass.; R. P. Sheehan, Ballou Lane, Williamstown, Mass.

Buenaventura (7)—E. C. Sternke, 320 Vista Del Mar, Camarillo, Calif.; Oliver La Plant, 325 N. "J" St., Oxnard, Calif.

Centre County (4)—W. L. Baker, 1184 Omeida St., State College, Pa.; W. J. Leiss, 1173 S. Atherton St., State College, Pa.

Charleston (3)—W. L. Schachte, 152 Grove St., Charleston 22, S. C.; Arthur Jonas, 21 Madden Dr., Dorchester Ter., Charleston Heights, S. C.

East Bay (7)—J. M. Rosenberg, 1134 Norwood Ave., Oakland 10, Calif.; C. W. Park, 6035 Chabolvn Ter., Oakland 18, Calif.

Erie (1)—R. S. Page, 1224 Idaho Ave., Erie 10, Pa.; R. H. Tuznik, 905 E. 25 St., Erie, Pa.

Fort Huachuca (7)—R. O. Burns, Electronic Prov. Gd., Ft. Huachuca, Ariz.; J. H. Homsy, Gen. Del., Warren, Ariz.

Lancaster (3)—G. W. Scott, Jr., Armstrong Cork Co., Lancaster, Pa.; G. E. Mandell, 522 E. King St., Lancaster, Pa.

Mid-Hudson (2)—E. A. Keller, Red Oaks Mill Rd., R.D. 2, Poughkeepsie, N. Y.; P. A. Bunyar, 760 South Rd., Poughkeepsie, N. Y.

Monmouth (2)—G. F. Senn, 81 Garden Rd., Little Silver, N. J.; C. A. Borgeson, 82 Garden Rd., Little Silver, N. J.

Northwest Florida (6)—K. L. Huntley, Mary Esther, Fla.; G. C. Jones, 12 N. Okaloosa Rd., Fort Walton Beach, Fla.

Orange Belt (7)—M. V. Kiebert, Jr., 1937 Elaine St., Pomona, Calif.; W. F. Meggers, Jr., 6844 De Anza Ave., Riverside, Calif.

Palo Alto (7)—W. W. Harman, Electronic Res. Lab., Stanford Univ., Stanford, Calif.; W. G. Abraham, 611 Hansen Way, c/o Varian Associates, Palo Alto, Calif.

Pasadena (7)—Officers to be elected.

Richland (7)—R. G. Clark, 1732 Howell, Richland, Wash.; R. E. Connally, 515 Cottonwood Dr., Richland, Wash.

Tucson (7)—R. C. Eddy, 5211 E. 20 St., Tucson, Ariz. (Chairman).

USAFIT (5)—W. T. Jones, USAFIT, Box 3125, MCLI, Wright-Patterson AFB, Ohio; J. J. Gallagher, Box 3482, USAFIT, Wright-Patterson AFB, Ohio.

Westchester County (2)—Joseph Reed, 52 Hillcrest Ave., New Rochelle, N. Y.; D. S. Kellogg, 9 Colonial Dr., Chappaqua, N. Y.

Wichita (6)—M. E. Dunlap, 548 S. Lorraine Ave., Wichita 16, Kan.; English Piper, 1838 S. Parkwood Lane, Wichita, Kan.

1955 INDUSTRIAL ELECTRONICS CONFERENCE

Sponsored by Detroit Section, PG on Industrial Electronics, and Michigan Section of AIEE

Detroit, Michigan, September 28–29

The Industrial Electronics Conference will be held at the Rackham Memorial Auditorium in Detroit, Michigan, September 28–29. The meeting is sponsored jointly by the Michigan Section of the American Institute of Electrical Engineers, the Professional Group on Industrial Electronics of the Institute of Radio Engineers and the Detroit Section of the IRE. Sixteen papers have been scheduled for the four technical sessions which will discuss automation, industrial measurement problems and new control system applications. The tentative program for the two-day conference is as follows:

Conference Registration will be $2.00 to members of the IRE and AIEE and $3.00 to all others. The conference hotel will be the Park-Shelton and reservations should be made prior to September 10. Information concerning advance conference registration and hotel reservations may be obtained from Guido Ferrara, 8106 West Nine Mile Road, Oak Park 37, Michigan.

Wednesday Morning
9:30—12:30

Session I—Electro-Optical Devices and Applications

"Miniature Strobelight System for a 60,000 rpm Bearing Tester," John Patraiko, Ford Motor Company.
"An Instrument to Count and Size Particles in a Gas," E. S. Gordon, Armour Research Foundation.
"Industrial Applications of a High Speed Spectrum Analyzer," N. L. Duncan, Raytheon Manufacturing Company.
"How Can Industry Use Television," H. F. Schneider, Radio Corporation of America.

12:30—2:30 p.m.

Luncheon at the Engineering Society of Detroit—$2.75 per person. Speaker: John R. Robertson, Chrysler Corporation.

2:30—5:30 p.m.

Session II—Measuring and Recording Instruments and Applications

"Capacitive Measurements of High Sensitivity and Their Applications to Industrial Testing and Control," George Revesz, Robertshaw Fulton Control Company.
"Some Applications of a Capacity Micrometer to High Speed, High Temperature Measurements," Ralph Condit, Ford Motor Company.
"Principles of Radioactive Gauging Applied to Measurements and Control in the Process Industry," D. C. Brunton, Isotope Products, Ltd.
"A Frequency-Modulated Magnetic Recorder," Walter Richter, Cutler-Hammer, Inc.

Thursday, September 29
9:30—12:30 p.m.

Session III—Process Control and Systems Analysis

"Automatic and Semi-Automatic Steel Flow Control Systems," Robert D. Morrow, Morrow Products, Inc.
"Problems in the Control of Nuclear Reactor Steam-Electric Power Plants," William Kerr, University of Michigan.
"Ultrasonic Impact Grinder—Industrial Tool," Kenneth W. Henderson, Raytheon Manufacturing Company.
"Automatic Controlled Electrolytic Grinding," Eugene Mittelmann.

12:30—2:30 p.m.

Luncheon at the Engineering Society of Detroit—$2.75 per person

2:30—5:30 p.m.

Session IV—Automation and Machine Tool Control

Title to be announced, Cledo Brunetti, General Mills, Inc.
"Automation Re-Examined," J. J. Graham, Radio Corporation of America.
"A Numerically-Controlled Cam Milling Machine," E. C. Johnson, Bendix Aviation Corporation.
"Two-Motion Duplicator for Machine Tool Control," A. J. Carr, Jr., Raytheon Manufacturing Company.

RADIO FALL MEETING PROGRAM

Hotel Syracuse,
Syracuse, N. Y.
October 17–19

Monday, October 17
9:30 A.M.

Report on the FCDA-RETMA Atomic Test of Commercial Equipment, R. H. Williamson, General Electric Co.

Additional paper to be announced.

2:00 P.M.

Sponsored by the IRE Professional Group on Reliability and Quality Control

Session Chairman, J. R. Steen, Sylvania Electric Co.

Type Test to Assure TV Performance Reliability, R. F. Rollman, C. Quirk, A. B. DuMont Laboratories.

Influence of Production Quality Distributions on Production Engineering, H. H. Mahuron, General Electric Co.

Two additional papers to be announced.

Tuesday, October 18

Sponsored by the IRE Professional Group on Broadcast and Television Receivers

9:00 A.M.

Transistorization Session

Session Chairman, W. P. Boothroyd, Philco Corp.

The Practical Application of Transistors in Monochrome Television Circuits, Ken James, Emerson Radio and Phonograph Corp.

Some Recent Advances in the Application of Transistors to R-F and I-F Radio Receiver Circuits, J. Karew, F. Mural, J. Tellier, Philco Corp.

Some Considerations of Transistor Video Amplifiers, M. C. Kidd, RCA.

A Discussion of the Design Problems Encountered in the Development of a Transistorized Radio Receiver, J. A. Worcester, General Electric Co.

Application of RCA Transistors to Battery-Powered Portable Receivers, John W. Englund, RCA.

2:00 P.M.

Television Session

Session Chairman, L. R. Fink, General Electric Co.

Methods of Measurement of Color Television Receiver Performance, Stephen P. Ronzheimer, Hazeltine Research Inc. and Richard J. Farber, Hazeltine Corp.

Generation of Television Sweep by Resonant Networks, Kurt Schlesinger, Motorola, Inc.

A Method of Measuring the Optical Sine-Wave Spectrum and Effective Bandwidth of TV Image Display Devices, O. H. Schade, Sr., RCA.

Design Considerations in the Reduction of Sweep Interference from Television Receivers, Alexander M. Intrator, General Electric Co.

Magnetic Field Effects on Color Receivers, Olaf H. Fernald, Westinghouse Electric Corp.

8:00 P.M.

Radio Fall Meeting Banquet

Toastmaster, J. D. Ryder, IRE President.

Speaker—to be announced.

Presentation of Radio Fall Meeting Plaque.

Wednesday, October 19
9:00 A.M.

Sponsored by the IRE Professional Group on Electron Devices

Session Chairman, R. R. Law, CBS Hytron.

High Frequency NPN Transistors, A. P. Kordalewski, General Electric Co.

Transistors for Portable Radios, Author to be determined, Texas Instrument Co.

Recent Improvements in the RCA-21AXP-22 Color Kinescope, R. B. Janes, L. B. Headrick, J. Evans, RCA.

22 Inch Rectangular CBS-Colortron, N. F. Fyler, P. Hambleton, T. Hodge, CBS-Hytron.

2:00 P.M.

Color and Brightness in Projected Pictures, R. M. Evans, Eastman Kodak Co.

SECOND ANNUAL MEETING OF PROFESSIONAL GROUP ON NUCLEAR SCIENCE

Sponsored by PG on Nuclear Science
September 14–16,
Oak Ridge, Tennessee

The Second Annual Meeting of the Professional Group on Nuclear Science will be held at the Center Theater in Oak Ridge, Tennessee, September 14–16. Registration fees for this meeting will be as follows: member PGNS—$2.50; member IRE, not PGNS—$4.00; non-member, $5.00. Registration and technical sessions at the Center Theater. A copy of the transactions will be sent to all persons registering for the meeting.

Officers of the meeting committee are: Chairman, Harold E. Walchli; Vice-Chairman-Treasurer, D. J. Knowles; Papers Committee, H. E. Banta; Arrangements Committee, R. W. Schede; Publicity Committee, E. Fairstein; Ex-officio, D. H. Loughridge.

A complete program, including abstracts, may be obtained from D. J. Knowles, Oak Ridge National Laboratory, P.O. Box P, X-10, Oak Ridge, Tenn.

TENTATIVE PROGRAM

Wednesday, September 14
10:00 A.M.–12:15 P.M.

Session I—Accelerators

Welcoming Address, R. W. Schede, Chairman, Oak Ridge Chapter PGNS.

History and Aims of the PGNS, M. A. Schultz, National PGNS Chairman.

The Brookhaven Electron Analogue, Ralph Kassner, Brookhaven National Laboratory.

The Microtron, a Nuclear and Electronic Research Instrument, H. F. Kaiser, Naval Research Laboratory.

Proton Beam Studies in a Fixed Frequency Cyclotron, Farno L. Green, Oak Ridge National Laboratory.

An Approximate Method for Obtaining the VSW on Cyclotron Dees, R. M. Donaldson, Oak Ridge National Laboratory.

2:00 P.M.–5:00 P.M.

Session II—Electronics

Phototube Voltage Regulators for Scintillation Counters, O. R. Harris and Bruce d'E. Flagge, University of Virginia.

Recent Advances in Modular Design of Electronics, W. G. James, A.C.F. Industries, Inc.

Multichannel Time Interval Analyzer, J. H. Neiler, H. E. Banta, W. M. Good, and E. C. Smith, Oak Ridge National Laboratory.

The 'Hard-Bottoming' Technique in Nuclear Instrumentation Circuit Design, C. C. Harris, Oak Ridge National Laboratory.

Four-Channel Counting System, D. W. Scott, Oak Ridge National Laboratory.

8:00 P.M.

Social and Smoker sponsored by Carbide and Carbon Chemicals Company, operators of Oak Ridge National Laboratory.

Thursday, September 15
9:00 A.M.–12:00 NOON

Session III—Reactor Controls and Pulse Height Analyzers

Electronic Analogue Devices for Design of Reactor Controls, E. R. Mann, Oak Ridge National Laboratory.

Control of a Two-Phase Reactor, John MacPhee, American Machine and Foundry, Atomics Division.

The Oak Ridge National Laboratory Serial Memory 120 Channel Pulse Height Analyzer, T. L. Emmer, Oak Ridge National Laboratory.

Circuits for Pulse Analysis, G. G. Kelley, Oak Ridge National Laboratory.

2:00 P.M.–5:00 P.M.

Session IV—Radiation Detection, Medical Instrumentation and Health Physics

Medical Radiation Instrumentation with Scintillation Spectrometers, P. R. Bell, Oak Ridge National Laboratory.

Radiation Fall Out Measurements, John Harley, AEC Operations New York.

Present Status of Halogen Quenched FM Tubes Using Transparent, Non-metallic, Electrically Conducting Cathodes—L. G. Clark, Sr., Naval Research Laboratory.

A Direct Current Integrator, F. M. Glass, Oak Ridge National Laboratory.

Instrument Requirements for Routine Medical Radioisotope Techniques, Theodore Fields, VA Hospital, Hines, Ill.

A Dual Function Gamma Monitor, R. E. Connally, General Electric Co., Hanford, Wash.

7:00 P.M.–10:00 P.M.

Banquet at the Oak Terrace, speaker to be announced.

Friday, September 16

Tour of unclassified facilities at Oak Ridge National Laboratory, Abbott Pharmaceutical Laboratory (radioactive drugs), and Museum of Atomic Energy.

Abstracts of Transactions of the IRE

The following issues of "Transactions" have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Broadcast and Television Receivers	Vol. BTR-1, No. 2	$ .95	$1.45	$2.85
Broadcast and Television Receivers	Vol. BTR-1, No. 3	$ .95	$1.45	$2.85
Circuit Theory	Vol. Ct-2, No. 2	$2.60	$3.90	$7.80
Electron Devices	Vol. Ed-2, No. 2	$2.10	$3.15	$6.30

* Public libraries and colleges may purchase copies at IRE Member rates.

BROADCAST AND TELEVISION RECEIVERS

Vol. BTR-1, No. 2, April, 1955

Automatic Gain Control of Transistor Amplifiers—W. F. Chow and A. P. Stern

Due to the dependence of transistor small signal parameters on the dc operating point, the gain of a transistor amplifier is function of the emitter current I_e and of the collector voltage V_c. An analysis of the variation of the series-parallel transistor parameters h_{ij} with the operating point shows that the gain decreases with decreasing I_e or decreasing V_c in the region of small values of I_e and V_c. The appreciable control power necessary to vary I_e or V_c (of possibly several amplifying stages) can be obtained advantageously by operating the controlled stages as dc amplifiers of the control signal and employing transistor detectors. Automatic gain control systems applicable to linear amplifiers and converters using I_e and V_c control have been developed, and their performance is described in this paper. Some problems arising in transistor agc systems are distortion, detuning and bandwidth variation.

Technical Requirements of the Australian Television System—A. J. McKenzie

Following a brief discussion of the important aspects of television standards, those used by various countries are described. The history of the Australian standards is then outlined and specific aspects of these standards are dealt with in some detail. The effect of introducing color television is considered. After summarizing the frequency channels available for Australian television services, standards of allocation of frequency and power are discussed. Finally some conclusions from tentative frequency allocation plans are given.

Preventing Fires from Electrical Causes in the Design and Manufacture of Radio and Television Receivers—H. T. Heaton

Minutes of the Meeting of the Administrative Committee of the IRE Professional Group on Broadcast and TV Receivers

Publication Committee Report

Notice for Papers for the Fall Meeting

Vol. BTR-1, No. 3, July, 1955

Selectivity and Transient Response Synthesis—R. W. Sonnenfeldt

The theory, design, and operation of a universal filter for the rapid synthesis of selectivity and transient responses are presented. This filter is useful in monochrome and color TV applications, and has the universality and ease of use normally associated with decade boxes. Like decade boxes, it enables the development engineer to design circuits that otherwise would require laborious and intricate calculations.

Low-pass, high-pass, band-pass, and band-stop characteristics, all at constant-time-delay, can be obtained in the basic operating range from 30 cps to 4.5 mc by throwing switches and adjustment of independent potentiometers. A simple theory, using elementary trigonometry, determines the settings. Transients can be synthesized by a series of small steps. This process produces the required phase and amplitude responses simultaneously. The settings in this case are determined by simple arithmetic. Experimental results are given for various selectivity curves and transients in the form of oscillograms and plotted curves.

A Transistor Sub-Carrier Generator for Color Receivers—L. J. Kabell and W. E. Evans

The era of transistorized television receivers on a commercial scale is still in the future. However, even at the present state of the art, laboratory tests on experimental transistorized color receivers have shown that certain of the circuit functions can be performed reliably enough with transistors to make them true competitors for vacuum tube circuits whenever the economic comparison becomes favorable. A representative circuit is the one used to generate the local 3.58 Mc sub-carrier.

A junction transistor circuit capable of performing the color synchronization function of a color television receiving system is described. The circuit employs a single high frequency junction transistor which serves as an oscillator-amplifier, phase detector, and current-controlled variable reactance in the generation of an accurately phased color reference carrier. A series mode quartz crystal filter in the feedback loop of the oscillator enables the circuit to perform well in the presence of noise interference up to a 1:1 signal-to-noise ratio.

The use of a transistor characteristic sometimes thought of as being objectionable illustrates the general principle of attempting to exploit those characteristics peculiar to transistors, rather than thinking of them as mere vacuum tube substitutes.

Differential Phase and Gain Measurements in Color Television Systems—H. P. Kelly

The presence of differential phase and gain distortion in systems used for the transmission of color television results in distortion of the colors being transmitted. A test set for measuring differential phase and gain is described. The set consists of two pieces of portable equipment, a transmitter and a receiver. Each has a self-contained power supply operated from 115 volt, 60-cycle power. The measurement is presented as a display on an oscilloscope. Scales of 0.5 db per inch differential gain and 2.5 degrees per inch differential phase are obtainable.

Operational Tests for Color Television—E. E. Gloystein

The advent of compatible color television has brought with it the need for several new types of test instruments to facilitate the adjustment of the specialized circuits required for generating, transmitting and receiving color television signals. This paper consists of descriptions of a selected group of test instruments which have been found particularly useful for routine operational tests in broadcasting plants and receiver service shops. Test generators which provide noise-free, artificial color signals are described, and techniques for using such signals for the adjustment of critical circuits in broadcast equipment and in home receivers are outlined briefly.

Light Amplification—P. E. Pashler

Three principal methods of light amplification are described. These are electron optical image tubes, contiguous layers of photoconductor and electroluminescent phosphor, and direct amplification of light in photoelectroluminescent film. Details of construction and relative merits of each are presented.

The Composite Video Signal—Waveforms and Spectra—J. B. Chatten, R. G. Clapp and D. G. Fink

Measured frequency spectra are given for the composite video signals of a variety of types of subject matter, in both color and monochrome. The subject matter includes colored flat fields, a bar chart and N.T.S.C. slides, as well as a monochrome flat field and resolution test chart. Data is presented in several forms, so as to show the relative amplitudes of all the main spectrum components, the general trends of energy distribution over the entire video frequency range, and the detailed fine structure showing the modulation sidebands on the individual spectrum components. The frequency interleaving of the luminance and chrominance components is clearly shown.

The waveforms resulting from scanning the various subjects are shown, including the synchronizing pulses and burst, and theoretical methods are discussed for predicting the resulting spectra.

The experimental methods which were employed are described.

This material will form part of a chapter of the forthcoming "Television Engineering Handbook," Donald G. Fink, Editor, to be published by McGraw-Hill.

UHF Tuner Local Oscillator Radiation—V. Mukai

Factors important to the design of a low radiation uhf television tuner have been empirically studied within the framework of the open field measuring technique specified by the FCC. Although there is some question as to the reliability of the open field method some significant results have been obtained. In particular it has been established that the magnitude of radiation is dependent on basic tuner circuitry as well as shielding and by-passing, namely on the type of oscillator, crystal circuit, and num-

ber of preselectors. An existing commercial model which employed a balanced oscillator and two preselectors was developed until radiation was below the proposed FCC limit of 500 uv/m. These developments were then successfully applied to a recently designed small package model. It was not necessary to make any sacrifice of overall tuner performance.

CIRCUIT THEORY

Vol. CT-2, No. 2, June, 1955

A Note on the Scattering Matrix of an Active Linear Two-Terminal Pair Network—J. E. Knausenberger

By separating the scattering matrix into matrix factors, a cascade of partial two-terminal pair networks is obtained, which represents the general nonreciprocal two-terminal pair network. Using a matrix factor to represent a source, the active circuit properties are separated from the passive ones and it is demonstrated that this source, which is simply related to the determinant can be moved to the terminal ends of the total network, where it may be combined with an external source.

The derived equivalent circuits do not employ gyrators but contain, besides the source mentioned, 2 two-terminal circuit elements, which characterize loss as well as inherent stability, and 2 "ideal transformers" rendering measures of amplitude and phase transfer.

An application to the transistor results in a novel equivalent circuit for that device.

Matrix Analysis of Oriented Graphs with Irreducible Feedback Loops—J. K. Percus

A generating function is obtained for the nonrepetitive closed loops of a network with unidirectional elements; from this is derived an analogous function for the irreducible loops. Criteria for the existence of loops are then established and the size of the smallest loop present determined; an asymptotic evaluation is made of the number of irreducible loops in a completely connected network. Further application of the generating function permits estimation of bounds for both irreducible and composite closed loops of a given order; less rigorous bounds are found by two perturbation techniques. The generating function is reformulated in terms of determinant-like quantities and application made to small networks.

Transformations preserving the irreducible loops of a system are discussed at length, following a delineation of the meaning of loop equivalence. Methods employed include elimination of branches, condensation of nodes, and decomposition of circuits, with criteria for their utilization being set forth. Finally, connection is made between the analysis of the present paper and those prevalent in more avowedly topological treatments.

Matrix Factorization—H. A. Schulke, Jr.

Minimum-Phase Transfer-Function Synthesis—R. H. Pantell

This paper presents a means for realizing a minimum-phase transfer-function $H(p)$ to within a constant multiplier, provided that $H(p)$ has no poles on the $j\omega$-axis including infinity. The network is in ladder form, without coupled coils, and every coil has an associated series resistance if $H(j\omega)\neq 0$ for any finite ω. Synthesis is illustrated for the specification of resistance termination at input and output, and for the specification of a resistor-capacitor termination at both ends. Regardless of the complexity of $H(p)$, the synthesis problem can always be reduced to the problem of synthesizing driving-point functions that are no more complicated than the ratio of quadratic polynomials in p. The method to be described is an extension of constant-resistance ladder network synthesis.

Neutralization and Unilateralization—C. C. Cheng

The subject of neutralization and unilateralization is of great interest in the field of transistor design because of the inherent bilateral property of transistors. This paper presents a systematic study of unilateralization in terms of generalized network theory presented in matrix form. Results are listed in tabulated form for easy use in practical circuit design. Examples illustrating the adaptation of the general procedure to the design of transistor amplifiers and vacuum-tube amplifiers are also included.

A General Matrix Factorization Method for Network Synthesis—E. C. Ho

This paper considers a new matrix factorization method for the synthesis of RLC two terminal-pair networks. Equivalent matrixes suitable for the synthesis of RLC ladder and parallel ladder networks are developed by linear transformations of matrix multiplication. The application of the method is demonstrated through the synthesis of a general minimum-phase transmission function as a ladder network and a general nonminimum-phase transmission function as a parallel ladder network. Ideal transformers are not required in the realized networks and superfluous elements in ladder networks are reduced considerably. Distribution of losses in reactive elements and arbitrary specification of resistive terminations are considered.

A Matrix Method for the Design of Relay Circuits—F. E. Hohn

This paper reviews the matrix method of synthesis of combinational multiple-output relay circuits first given in reference 44. Further examples are provided and it is shown how the method may be applied to sequential circuits as well. The purpose of presenting this material in the present connection is to illustrate the use of unconventional techniques with unconventional types of matrixes in switching circuit design.

Generalized Mesh and Mode Systems of Equations—M. B. Reed

Based on the obvious complexity of present-day problems to be solved and the increasing availability of computers, the broadest possible base for establishing the differential equations of an electrical network is an urgent need. This paper presents, on a topology, Laplace transform base, the differential equations and their solutions for any network describable in terms of linear differential equations with constant coefficients. These systems of equations by successive matrix partitioning and change of variable, lead to generalized "mesh" and "node" systems of unrestricted character.

Use of Tchebycheff Functions in Dealing with Iterated Networks—H. L. Armstrong

Expressions for the transmission matrices of four-terminal networks in terms of Tchebycheff functions have been given previously. Here, relations among these functions are used to give a convenient formula for the voltage gain of such networks; the transfer impedance etc., could be handled similarly. As an example, the low-pass network formed by cascading inductance-capacitance T sections is discussed.

Regeneration Analysis of Junction Transistor Multivibrators—D. O. Pederson

The two-transistor, collector-coupled relaxation circuit is a prototype configuration from which various switching circuits such as monostable multivibrators and flip-flops can be derived. Although the operational possibilities derived from the basic configuration, e.g., astable, monostable, or bistable, may differ in the details of triggering and stable point operation, the circuits of this class will have in common the key factor of the regenerative switching behavior. In addition, the circuits will have similar transient behavior immediately following the regeneration, having to do with the drive of one transistor off and the other into saturation.

The operation of the transistors for a complete cycle of operation can be divided into regions, each region of which can be characterized by an approximate linear equivalent circuit. Hence, a piece-wise linear analysis can be made of the operation of these circuits. In this paper, attention will be centered on the analysis of the regenerative switching mechanism which occurs when the transistors are in the active region. A major result of the analysis is the derivation of a simple formula for the switching time. The analysis also provides a fundamental inequality which must be satisfied in order to obtain regenerative switching. An extension of the results leads to an expression for the maximum repetition frequency.

In setting up the regeneration analysis, elementary design data are established. From this data, a minimum value for α_0 can be specified if sharp, rectangular output waveforms are desired.

Predictions Based on Maximum Oscillator Frequency—P. R. Drouilhet, Jr.

Considerable difficulty is encountered in directly measuring the parameters of a transistor at very high frequencies. An approximate high-frequency equivalent circuit for a transistor is presented, and several techniques for measuring the alpha-cutoff frequency are discussed. An indirect technique is presented involving the measurement of the maximum frequency at which the transistor can oscillate, and it is shown that this leads to a simple and accurate determination of the alpha-cutoff frequency. This maximum frequency of oscillation can also be used to predict the approximate gain obtainable from a transistor at high frequencies, and the efficiency which may be realized from the transistor used as a high frequency oscillator.

Frequency Response of Theoretical Models of Junction Transistors—R. L. Pritchard

For a grown-junction transistor, the concept of a constant base-spreading resistance may not be valid at high frequencies, owing to the distributed nature of the transistor parameters in the transverse direction of the base. However, results of a theoretical analysis of an appropriate two-dimensional model have shown that this type of transistor may be represented by the same type of model as that normally used for the fused-junction transistor, but with the constant base spreading resistance of the latter model replaced by a *complex* frequency-dependent *base impedance*. These two types of models represent limiting cases which should be useful for calculating circuit performance of practical junction transistors. In this paper, a method of comparing circuit performance of these two types of transistor models is described for both grounded-base and grounded-emitter configuration, using the series-parallel, or h, parameters. Under simplifying conditions, either type of transistor model in either configuration can be described by three normalized functions of frequency relative to α-cutoff frequency plus three additional constants. Simple relations are shown to exist between grounded-base and grounded-emitter parameters. Polynomial representations are given for the h parameters for both grounded-base and grounded-emitter operation, and simplified equivalent circuits are presented. To illustrate this method of circuit analysis, numerical examples are given for power gain and input resistance for a one-stage amplifier terminated in a pure resistance. Finally, the subject of maximum available power gain also is discussed briefly.

Constant-Resistance AGC Attenuator for Transistor Amplifiers—C. R. Hurtig

The gain of a wideband tuned transistor amplifier may be varied over a large range by an external control voltage, accompanied by

only slight changes in amplifier bandwidth or center frequency, by means of a unique constant-resistance *agc* attenuator. The control power required by this attenuator is in the milliwatt region. At audio frequencies this relatively simple attenuator may be used to obtain the arithmetic operations of multiplication and division.

Weighted Least-Squares Smoothing Filters—L. A. Ule

This paper differs from many others on least square filtering, in that no explicit note has been taken of the noise spectrum, at least no more than is taken when one fits a curve by least squares to a set of data. The concept of a least weighted error, not new in curve fitting, has, to the author's knowledge, never before been applied to a filter—other workers have uniformly weighted their signal over a fixed interval of length T, with the result that the filters so derived cannot be realized with lumped constants. If this artificial constraint is removed as it is here, lumped constant filters are possible. The expression for the filter weighting function is obtained with a bare minimum of elementary mathematics; a very slight generalization leads to an expression for a time-varying filter weighting function when this is required. The nonlinear least-square filter is considered but no general solution is given.

The paper is replete with examples and is directed to the average engineer. Although some original material is presented, a large part of the paper may be considered tutorial.

Note on a Logarithmic Approximation for Use with Singularity Plots—H. E. Tompkins

The analysis or synthesis of band-pass networks using plots of the singularities of their transfer functions can often be shortened and simplified by using a suitable logarithmic transformation of the complex-frequency s-plane which permits a simple approximate calculation of the transfer function. This approximation is good for over-all-bandwidth ratios of 2 to 1 or less as compared with a usable bandwidth ratio of 1.2 to 1 for the conventional narrow-band approximation. This transformation is not intended for use with an electrolytic tank, for which better methods have been described in the literature. It does not have the power of certain conformal transformations, but is considerably simpler. Unlike the wide-band low-pass to band-pass transformation it is not limited to pole and zero patterns of particular symmetry. This approximation also has interesting properties as an aid in the factoring of network polynomials.

Circuits with Quantized Feedback—Rajko Tomovitch

The paper deals with a special class of feedback circuits in which the feedback path is closed at a discrete set of time instants that depend upon two arbitrary inputs $g(t)$ and $v(t)$. Since the feedback signal is quantized in magnitude as well as in time, these circuits possess novel properties not found in ordinary linear feedback systems. The equations of such circuits are established and it is shown that these circuits may perform various mathematical operations, such as yielding inverse or reciprocal functions. Application to a communication method similar to delta modulation is described.

Reviews of Current Literature

"The Complete Specification of a Network by a Single Parameter"—M. S. Corrington, T. Murakami and R. W. Sonnerfeldt . . . Reviewed by *A. D. Perry*

"Extension de la méthode du diagramme de phase généralizé dans l'étude de la stabilité des systèmes linéaires"—P. Lefevre . . . Reviewed by *V. Belevitch*

"A New Method of Synthesis of Reactance Networks"—A. Talbot . . . Reviewed by *H. J. Orchard*

"Synthesis of Distributed Amplifiers for Prescribed Amplitude Response"—A. D. Moore . . . Reviewed by *W. H. Kautz*

Correspondence

Comment on B. J. Bennett's Paper, "Synthesis of Electric Filters with Arbitrary Phase Characteristics" . . . D. Helman

Reply to Mr. Helman's Letter . . . B. J. Bennett

"The Spelling of the Name Napier" . . . H. A. Wheeler

News

PGCT News

ELECTRON DEVICES

Vol. ED-2, No. 2, April, 1955

Plasma Frequency Reduction Factors in Electron Beams—G. M. Branch and T. G. Mihran

The electron plasma frequency reduction factor has proved to be a fundamental design parameter in all types of microwave tubes employing long electron beams. This factor is here calculated for a variety of beam shapes and drift-tube cross sections, and the results are presented in a series of graphs. One interesting result is that the reduction factor for an annular beam depends primarily on the width of the annulus and is relatively independent of the radius of curvature of the beam.

Noise in Traveling-Wave Tubes—A. G. Mungall

A number of experimental traveling-wave tubes have been built for operation in the 3.2 centimeter wavelength region, and series of noise measurements have been made on these tubes. The periodic dependence of noise figure on the separation between the electron gun and the circuit entrance of the traveling-wave tube has been investigated and various characteristics of these curves have been discussed. A modified noise theory has been suggested; comparisons between it and the experimental results show fair agreement. Noise figure reduction by the use of a triode gun has been investigated for two of the experimental tubes, and a method of analysis that should lead to theoretical justification is suggested.

Writing Speed and Tonal Range of Dark Trace Tubes—Seymour Nozick

The writing speed and tonal range properties of dark trace tubes are analyzed and figures of merit are outlined. Experimental results are presented. Writing speed of a dark trace tube varies linearly with the ratio of beam current to spot size. The information writing rate varies directly with the beam current and inversely with the square of the spot size. The tonal range of dark trace tubes varies directly with the ratio of maximum contrast to spot size.

Suppression of Backward-Wave Oscillation by Filter Helix Methods—A. E. Siegman and H. R. Johnson

An experimental study has been made of the filter helix properties of a periodically loaded helix, using a special traveling-wave tube. The filter helix is shown to possess filter-like frequency pass bands and stop bands and a phase velocity characteristic such that forward- and backward-wave phase velocities are separated, making the filter helix a useful circuit for traveling-wave amplification at high ka (ratio of helix circumference to free space wavelength) without danger of backward-wave oscillation. Certain difficulties associated with backward-wave oscillation frequency pushing were found to arise in filter helices, but useful gain was obtained at ka greater than 0.5 using filter helix techniques.

A Wide-Band Square-Law Circuit Element—A. S. Soltes

A square-law circuit element with operating frequency range from zero into the vhf region is described. Its dynamic range and accuracy capabilities vary with the particular conditions under which it is operated; accuracies within less than one per cent of full scale and output dynamic ranges of over 100 db have been achieved. Frequency response limitations and possible sources of error are analyzed. Experimentally determined characteristics are presented and noise properties, dynamic range, and accuracy potentialities evaluated.

Low-Frequency Circuit Theory of the Double-Base Diode—J. J. Suran

The double-base diode is a single-junction semiconductor triode. When an electric potential is applied between the two ohmic contacts, a negative-resistance is obtained between the junction and one of the ohmic contacts. This negative resistance is bounded by two positive-resistance regions, one of considerably high magnitude which corresponds to the cut-off state and one of very low magnitude which corresponds to a saturating condition. The magnitude of the negative resistance is related to the ratio of majority-to-minority carrier mobilities. Small-signal low-frequency equivalent circuits are developed to approximate the double-base diode in each of the operating regions of the negative-resistance characteristic and equations for current and voltage amplification, input and output resistance and power gain are developed. The important circuit parameters are related to the physical constants of the device.

Space-Charge Conditions in a Reflected Flow of Electrons—J. T. Wallmark

Some characteristic features of a reflected flow of electrons are described, in particular variations of the virtual cathode and transit time with respect to current. This has been accomplished by finding new solutions to well-known basic equations treated earlier by Fay, Samuel, Shockley, Salzberg, Haeff and others. The results are applicable to problems where the current is varied while earlier solutions were considering the potential as variable. The theoretical results are found to be in agreement with experimental results obtained on reflex klystrons and space-charge deflection tubes.

1955 STUDENT AWARDS

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January 1—July 29, 1955

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Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Index to the Abstracts and References published in the PROC. IRE from February, 1954 through January, 1955 is published by the PROC. IRE, April, 1955, Part II. It is also published by *Wireless Engineer* and included in the March, 1955 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

543.2 **2174**

The Propagation of Sound Pulses along Metal Bars—W. Guth. (*Acustica*, vol. 5, no. 1, pp. 35–43; 1955. In German.) The stationary-phase method is applied in developing the theory of pulse propagation along a bar. Experiments using schlieren technique were made on cylindrical steel bars immersed in water, with an electric spark as pulse source. Agreement of results with theory was satisfactory. In the case of thick bars, a wave of higher order was observed, presumably the first-order skew-symmetric wave.

534.213 **2175**

Sound Propagation in the Atmosphere and Audibility of Warning Signals in Ambient Noise—P. Baron. (*Ann. Telecommun.*, vol. 9, pp. 258–274; October, 1954.) An account of experiments carried out in 1938–1939 in the Vals d'Yonne district and in Paris. Meteorological and sound-pressure measurements were made. Raising the sound source from ground level to 37 m produced a marked increase of received sound power for distances <1 km. A mean curve of propagation was derived from the experimental data. The directional effect of the wind makes it desirable to have several sound sources to cover a given area. The degree of audibility satisfactory in the presence of noise was determined in a series of laboratory experiments. Computation shows that two sources having a combined power equal to that of a single source have a greater range.

534.64:534.2 **2176**

Tubes and Resonators. Computation and Measurement of Some Acoustic Resistances—J. Guittard. (*Acustica*, vol. 5, no. 1, pp. 7–18; 1955. In French.) Assuming that the linear dimensions of the enclosed space are small compared with $\lambda/4$, where λ is the wavelength of the sound, acoustic resistance is computed from elementary theory for a circular aperture in a thin wall, for a tube opening on to (a) a semi-infinite and (b) an infinite space, and for a discontinuity of tube cross section. Satisfactory agreement between theory and experiment is found.

534.78:621.39 **2177**

The Effect of Severe Amplitude Limitation on Certain Types of Random Signal: a Clue to the Intelligibility of "Infinitely" Clipped Speech—(*See* 2412.)

534.833 **2178**

Subjective Assessment of Sound Insulation, using Electrical Simulation of Transmission-Loss Curves—H. J. Rademacher. (*Acustica*, vol. 5, no. 1, pp. 19–27; 1955. In German.)

534.85 **2179**

Frequency Irregularity in Rooms—A. F. B. Nickson and R. W. Muncey. (*Acustica*, vol. 5, no. 1, pp. 44–47; 1955.) The sounds of the speaking or singing voice and of musical instruments are often modulated in frequency and/or contain transients. These are the sounds of real importance in assessing the acoustic characteristics of auditoria. Frequency irregularity measurements using only pure tones may therefore give misleading results. Experiments with tones modulated in frequency at 6 cps over a range of ± 3 per cent show that the frequency-irregularity curves for such tones are much smoother than the corresponding ones for pure tones.

534.86:621.396.712.3 **2180**

On the Structural and Room Acoustics of the Multipurpose Studio Unit at Broadcasting House, Hamburg—G. Venzke. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 6, nos. 11/12, pp. 229–236; 1954.) The two-story building contains two groups of studios designed principally for recording plays. These comprise two "dead" studios, two studios with normal reverberation time, two studios in which the reverberation time can be changed, and one underground reverberation room. Reverberation-time/frequency curves are shown and results of measurements of the sound insulation achieved by the double-leaf, floating-floor construction of the studios are noted.

621.395.6:621.372.5 **2181**

Equivalent Quadripole Networks for Electromechanical Transducers: Part 1—W. Reichardt and A. Lenk. (*Acustica*, vol. 5, no. 1, pp. 1–6; 1955. In German.) Es, em and electrodynamic transducers are considered, with particular reference to the frequency-independent transformer or gyrator used as the coupling element between the electric and mechanical elements.

621.395.61+621.395.623.7].012.12 **2182**

Directivity Characteristics of Electroacoustic Transducers—A. C. Raes. (*Ann. Télécommun.*, vol. 9, pp. 313–314; November, 1954.) A method of obtaining the polar characteristics of loudspeakers and microphones is described which does not require the use of an anechoic chamber. The transducer under test and the recording instrument are arranged at distances from the reflecting wall such that the characteristic can be recorded during the period preceding the arrival of the reflected sound.

621.395.623.8 **2183**

Sound System for Plenary Hall of United Nations General Assembly Building—C. W. Goyder and L. L. Beranek. [PROC. IRE, (*Australia*), vol. 16, pp. 38–44; February, 1955. *1953 IRE Convention Record*, part 3, pp. 26–34; 1953.] This specially designed sound-reinforcement system has a linear frequency response over the range 300 cps–6 kc. Time-delay systems are used to improve intelligibility. Measured performance characteristics are presented.

621.395.625.3:621.397.6:778.5 **2184**

Methods of Picture-Synchronized Sound Recording in Television—Gondesen. (*See* 2450.)

ANTENNAS AND TRANSMISSION LINES

621.372.2 **2185**

Theory of the Harms-Goubau Wire Waveguide at Metre Wavelengths—G. Piefke. (*Arch. elekt. Übertragung*, vol. 9, pp. 81–93; February, 1955.) Whereas at centimeter and decimeter wavelengths thin dielectric layers provide adequate concentration of the wave energy, at meter wavelengths, and particularly for dealing with bends, thick dielectric layers are required. Theory is developed for this case, and is extended to include the use of a separate dielectric tube shielding the line from precipitation. Both the leakage attenuation and the resistive attenuation increase with rising frequency, the former increasing also as dielectric constant decreases. For a given degree of energy concentration, the attenuation has a minimum value for a particular value of wire radius, but this value may be unacceptable for economic reasons. Attenuation due to a coating of ice on the line is investigated. Neglecting icing effects, a relay section of length 15 km with a loss of 5n at 1 mλ is practicable.

621.372.2+621.372.54]:512.831 **2186**

The Matrix Approach to Filters and Transmission Lines—Fisher. (*See* 2231.)

621.372.21:621.396.67 **2187**

The Behaviour of the Open End of a Coaxial Line—V. Pfirrmann. (*Arch. elekt. Übertragung*, vol. 9, pp. 8–12; January, 1955.) A theoretical study is made of the field pattern and variations at the junction of a coaxial line and a circular waveguide; the method of least squares is used. Particular cases treated are (a) frequency

below the cut-off value for the waveguide, and (b) radiation conditions, the waveguide being excited in the E_{01} mode.

621.372.221:621.395.97 2188

Programme Circuits on Cable Pairs Loaded at 500-Yard Intervals—J. L. W. Morgan and W. S. Ash (*P.O. Elec. Eng. Jour.*, vol. 47, part 4, pp. 193–196; January, 1955.) Cable links between telephone exchanges which will be suitable for normal traffic when not required for BBC program circuits for outside broadcasts are obtained by loading standard cable pairs with 22-mH coils at 500-yard nominal spacing. Sufficient tolerance on the spacing to permit existing jointing chambers to be used is achieved by artificially increasing the mutual capacitance of the circuits.

621.372.8 2189

Propagation of a Signal in a Waveguide—P. Poincelot. (*Ann. Télécommun.*, vol. 9, pp. 315–317; November, 1954.) An approximate formula is derived for the amplitude of the signal at a given time. Its application is illustrated by a numerical example for a guide of length 3 km and radius 3.825 cm. The duration of the initial transient state is found to be proportional to the square root of the length.

621.372.8 2190

Stability of the H_{01} Mode in Circular Waveguides, and the Occurrence of Harmonic Modes on Deformation to an Elliptical Cylinder—J. Kornfeld. (*Arch. elekt. Übertragung*, vol. 9, pp. 29–38; January, 1955.) Analysis involving use of Mathieu functions is presented. A numerical example indicates that a 10 per cent deformation gives rise to an H_{11} wave whose amplitude is about 0.15 that of the original H_{01} wave.

621.372.8:538.614 2191

Propagation of Electromagnetic Waves in an Anisotropic Gyromagnetic Medium in a Rectangular Waveguide—A. Chevalier, T. Kahan and E. Polacco. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1323–1324; March 21, 1955.] Analysis is presented for the case where the gyromagnetic medium is made anisotropic by application of a magnetic field perpendicular to the direction of propagation. General expressions are derived for the three electric-field components.

621.372.8:621.318.134 2192

Temperature Dependence of the Microwave Properties of Ferrites in Waveguide—B. J. Duncan and L. Swern. (Proc. IRE, vol. 43, pp. 623–624; May, 1955.) Report of an experimental investigation on a rod of ferramic *A*-106 in a waveguide of internal diameter 0.937 inch, with a static magnetic field parallel to the direction of propagation. The results indicate that at some temperature between 23 degrees and 500 degrees C. the ferrite absorption loss for this specimen is practically independent of frequency.

621.396.67+621.397.62 2193

New Television Receivers and Aerials [in Western Germany]—W. W. Diefenbach. [*Funk-Technik* (*Berlin*), vol. 10, pp. 88–91; February, 1955.] A brief survey of commercially available equipment.

621.396.67 2194

Scattering of Electromagnetic Waves by Wires and Plates—D. B. Brick and J. Weber. (Proc. IRE, vol. 43, p. 628; May, 1955.) Comment on 930 of May and author's reply.

621.396.67 2195

Method of Measurement of Aerial Gain by Electrical Integration of the Characteristic Function—J. Munier. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1411–1413; March 28, 1955.] An arrangement is described which operates in conjunction with a polar-diagram recorder with square-law receiver to give a direct reading of the gain for antennas with rotationally symmetrical characteristics. The af output of the receiver is applied to the primary of a rotating transformer mechanically coupled to the rotating antenna support; the secondary voltage is detected by means of a rectifier whose time constant is large compared with the af period.

621.396.67:621.372.2 2196

Gap Problem in Antenna Theory—R. King. (*Jour. Appl. Phys.*, vol. 26, pp. 317–321; March, 1955.) Discussion indicates that the so-called gap problem [see e.g. 24 of 1948 (Infeld)] is more properly to be considered as a problem of the coupling between the transmission line and the antenna.

621.396.67:621.396.93 2197

Aerials for Mobile Radio Services—W. Stohr and D. Bassler. (*Frequenz*, vol. 8, pp. 357–368; December, 1954.) A survey covering wide-band antennas comprising full-wave dipoles, corner reflectors, collinear antennas, Yagi arrays, multichannel $\lambda/4$ radiators and folded dipoles, all for meter wavelengths.

621.396.67.012 2198

Wave Launching from a Conical Aerial—V. Pfirrmann. (*Arch. elekt. Übertragung*, vol. 9, pp. 98–101; February, 1955.) The near field of a $\lambda/4$ conical antenna is computed, and the wave-launching mechanism is illustrated by field patterns.

621.396.67.095.12 2199

Elliptic Polarization of Electromagnetic Radiation—M. Bouix. (*Ann. Télécommun.*, vol. 9, pp. 275–281, 298–304 and 345–351; October–December, 1954.) Fundamental formulas are developed in detail, and the general expressions of Kottler and Goudet are adapted for elliptic polarization. The results are used to obtain formulas for the gain and effective surface area of antenna systems.

621.396.674.1:621.318.134 2200

Ferrite-Cored Antennae—C. A. Grimmett. [*Proc. IRE* (*Australia*), vol. 16, pp. 31–35; February, 1955. Discussion, pp. 36–37; *1954 IRE Convention Record*, Part 7, pp. 3–7.] A survey of developments in the use of this type of antenna for reception of broadcast at 540 kc–1.6 mc. The influence of length, diameter, winding and core material on sensitivity is discussed; ferrite-cored and air-cored antennas are compared.

621.396.674.3:621.372 2201

The Radiation of a Hertzian Dipole over a Coated Conductor—D. B. Brick. (*Proc. IEE*, Part C, vol. 102, pp. 104–121; March, 1955; Digest, *ibid.*, Part B, vol. 102, pp. 392–395; May, 1955.) The study presented was undertaken in an effort to explain certain characteristics of an antenna field pattern measured over an aluminium ground screen presumed to have a thin coating of aluminium oxide. Analysis is given for the case where the dipole is (a) above the dielectric coating, (b) in the dielectric coating, and (c) lying in the conductor/dielectric interface. Values of the field potentials are obtained for both electric and magnetic dipoles. The power from the dipole is partly radiated and partly surface-guided; numerical values for the ratio between the two are derived for some particular cases.

621.396.676 2202

Aircraft Antennas—J. V. N. Granger and J. T. Bolljahn. (Proc. IRE, vol. 43, pp. 533–550; May, 1955.) A survey of antennas for communication and navigation on conventional aircraft.

621.396.677.45 2203

Radiation Characteristics of a Conical Helix of Low Pitch Angle—J. S. Chatterjee. (*Jour. Appl. Phys.*, vol. 26, pp. 331–335; March, 1955.) Experiments are reported on a conical helix with a pitch angle much smaller than in the previous investigation (2899 of 1953) and with a larger ground plane. Mounted with the apex at the bottom and a short distance above the ground plane, the arrangement provided radiation in the axial direction over a frequency range of 100–500 mc. Radiation pattern and current distribution along the helix were determined experimentally; the radiation pattern is also determined by computation, assuming a linear current distribution.

621.396.677.85 2204

Virtual Source Luneberg Lenses—G. D. M. Peeler, K. S. Kelleher and H. P. Coleman. (*Trans. IRE*, vol. AP-2, pp. 94–99; July, 1954.) An investigation is made of lenses formed by portions of a spherical Luneberg lens bounded by a pair of plane reflectors intersecting along a diameter; wedge angles up to 180 degrees are considered. Multiple virtual sources corresponding to the reflections of the feed point can be obtained. Calculations are made of the positions of these virtual sources and the width of the corresponding beam. Methods of eliminating undesired beams include the placing of absorptive material at the edges of the lens. Data obtained experimentally on a two-dimensional lens with wedge angle of 180 degrees are in good agreement with calculated values.

AUTOMATIC COMPUTERS

681.142 2205

Computer for Universal Application—(*Elec. Times*, vol. 127, pp. 319–320; February 24, 1955.) Brief description of DEUCE, a commercially available computer developed from the ACE; it has punched-card input and output systems, mercury delay lines for short-term storage and a magnetic recording drum for long-term storage. The total floor space occupied by the equipment is 14 feet by 4 feet 6 inches.

681.142 2206

An Outline of an Electronic Arithmetic Unit—W. Woods-Hill. (*Electronic Eng.*, vol. 27, pp. 212–217; May, 1955.) Discussion of the design of a digital-computer unit capable of performing and checking the operations represented by $A\times B\pm D=C$.

681.142 2207

Pulse-Switching Circuits using Magnetic Cores—M. Karnaugh. (Proc. IRE, vol. 43, pp. 570–584; May, 1955.) Design theory is presented for nonstorage uses of cores with rectangular hysteresis loops in digital computers. To facilitate determination of the sense of the em processes a method of representation due to Mayer is used in which the magnetic-circuit elements are replaced by elements of a mirror system. Devices for eliminating output voltage on "shuttling" the core are discussed. Operating frequencies >100 kc are attainable. 31 references.

681.142 2208

A Statistical Method for Solution of the Laplace Differential Equation using Electronic Computers—H. Harmuth. (*Acta Phys. austriaca*, vol. 9, pp. 27–32; December, 1954.) An electrical analog of the Galton board is used in which the rolling balls are replaced by pulses and the pins by pulse-storage units.

681.142:621.37:535.32 2209

Analogue Machine for Calculation of the Complex [refractive] Index of a Body from its Reflection Coefficient—M. Hénon. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1305–1306; March 21, 1955.] A resistance network is discussed by means of which the phase shift of light reflected by a body can be calculated if the reflection coefficient is known for all wavelengths; the refractive index can hence be determined.

681.142:621.385.832 2210
The Function of Basic Elements in Digital Systems—C. B. Speedy. (*Proc. IEE*, Part C, vol. 102, pp. 49–56; March, 1955. Digest, *ibid.*, Part II, vol. 101, pp. 677–679; December, 1954.) The three basic elements of digital computers are (a) a bistable element for storing a digit, (b) a gate for controlling the passage of a digit, (c) a diode for controlling the direction of passage. A unit in the form of a beam deflection tube has been designed including a bistable element together with two gating elements. *See* also 2485 below.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.21.012 2211
The Geometric Transformation of Impedance Diagrams—A. Železnikar. (*Telefunken Ztg.*, vol. 27, pp. 252–253; December, 1954.) An extension of the work of Briner and Graffunder (2576 of 1953) to the general case. For a correction to the earlier paper see *ibid.*, vol. 27, p. 254; December, 1954.

621.3.042:621.396.822 2212
Barkhausen Noise from a Cylindrical Core—D. Haneman. (*Jour. Appl. Phys.*, vol. 26, pp. 355–356; March, 1955.) The method used by Krumhansl and Beyer for calculating Barkhausen noise (298 of 1950) is simplified by considering an exciting field with a sawtooth rather than a sinusoidal waveform.

621.3.066.6 2213
The Variation with Current and Inductance of Metal Transfer between Platinum Contacts—J. Riddlestone. (*Proc. IEE*, Part C, vol. 102, pp. 29–34; March, 1955.) Continuation of investigation described previously [3495 of 1953 (Warham)]. Curves of the net transfer are given for currents and inductances in the ranges 1.8–7.6 a and 0.06–117 μh respectively. Four different types of transfer, termed "bridge," "short arc," "long arc," and "reversed short arc," may be involved. The results indicate that the life of Pt contacts could be improved by controlling the effective circuit inductance at break to about 0.6 μh.

621.314.223 2214
Analytical Approach to the Variable Turns-Ratio Autotransformers—E. Mishkin. (*Trans. Amer. IEE*, Part III, *Power Apparatus and Systems*, vol. 72, pp. 669–673; August, 1953.)

621.314.7:621.375.4 2215
High-Frequency Amplification using Transistors—E. Kettel. (*Telefunken Ztg*, vol. 27, pp. 245–251; December, 1954.) Measurements were made on transistors Type *OC 601* and *OC 602* in the grounded-base connection to assess them for use in narrow-band IF amplifiers. The voltage feedback effect may lead to instability, hence neutralization is essential. At the higher frequencies (~500 kc) the stage gain is reduced mainly by the base resistance and the collector barrier capacitance.

621.314.7.012.8 2216
Theory of Equivalent Circuits for Junction Transistors—L. Oertel. (*Telefunken Ztg*, vol. 27, pp. 230–237; December, 1954.) An equivalent circuit is discussed which differs from that of Pritchard (2537 of 1954) only in respect of the arrangement of the current and voltage sources. A simplified equivalent circuit including a triode tube is also presented.

621.314.7.012.8 2217
The Frequency Dependence of [junction-] Transistor Quadripole Parameters—E. Kettel and G. Meyer-Brötz. (*Telefunken Ztg.* vol. 27, pp. 237–245; December, 1954.) The grounded-base connection only is considered. Oertel's equivalent tube circuit (2216 above) is a satisfactory approximation up to a frequency about half that of α cut off, but a proper description of rf performance requires the addition of a base resistance and leakage conductance to the basic equivalent circuit. Measurements on a Type-*OC 601* transistor show satisfactory agreement with theory.

621.318.4+621.314.2]:621.3.015.3 2218
Field Theory of Wave Propagation along Coils—H. Poritsky, P. A. Abetti and R. P. Jerrard. (*Trans. Amer. IEE*, Part III, *Power Apparatus and Systems*, vol. 72, pp. 930–938; October, 1953. Discussion, pp. 938–939.) Expressions are derived for phase and group velocites and surge impedance in air-cored and iron-cored coils and transformer windings. Theoretical and experimental data for frequencies up to 600 kc are compared.

621.318.42 2219
The Design of Coils for the Production of High Magnetic Fields—A. N. Ince. (*Proc. IEE*, Part C, vol. 102, pp. 25–28; March, 1955. Digest, *ibid.*, Part A, vol. 102, p. 100; February, 1955.) Design curves are presented for coils of rectangular cross section for producing intense transient magnetic fields, using energy obtained from a bank of charged capacitors.

621.318.57 2220
A Reversible Binary Counter—R. W. Fenemore. (*Elect. Eng.*, vol. 27, pp. 204–206; May, 1955.) A multivibrator-type counter performs addition or subtraction depending on application of a control signal to a two-gate system interposed between consecutive counter stages. The control signal may be obtained from a multivibrator circuit similar to the counter stages. Applications in digital-analog conversion and in an interpolator are discussed.

621.318.57:621.314.7 2221
Transistor Choppers for Stable DC Amplifiers—R. L. Bright and A. P. Kruper. (*Electronics*, vol. 28, pp. 135–137; April, 1955.) Two fused-junction transistors driven at power frequency are used in a switching circuit for converting weak dc input signals into proportional square wave ac signals.

621.318.57:621.38 2222
Multi-electrode Counting Tubes—K. Kandiah and D. W. Chambers. (*Jour. Brit. IRE*, vol. 15, pp. 221–232; April, 1955. Discussion p. 232.) Applications other than straightforward counting operations are discussed for decimal counting tubes of various types. The design of a pulse amplitude analyzer using trochotrons and dekatrons in a matrix system is outlined. The life of the tubes is comparable to that of ordinary tubes.

621.37.+621.396.621]049.75 2223
Investigations of Laboratory Production of Printed Circuits for Communication Equipment—W. Götze. (*Fernmeldetech. Z.*, vol. 8, pp. 83–88; February, 1955.) A brief report is presented on the practical aspects of very-small-scale production of printed circuits, using the simplest tools only. The electrical and mechanical properties of 12 different chassis materials are tabulated as well as the properties and the treatments required by seven different conducting materials. Applications described include the production of components, an amplifier, an *RC* oscillator and a heterodyne receiver for the frequency band 500 kc–15 mc; this receiver contains no inductive components (see also 958 of 1953).

621.372.412 2224
Piezoelectrically-Activated Low-Frequency Mechanical Resonators—J. F. W. Bell. (*Jour. Sci. Instr.*, vol. 32, pp. 52–54; February, 1955.) Discussion of resonators comprising plates or bars carrying two relatively small piezoelectric crystals. The parameters of the equivalent electrical quadripole are determined for flexural modes of some circular Chladni's plate type resonators. Application to stabilization of sinusoidal oscillators is briefly described.

621.372.5 2225
On the Physical Realizability of Linear Non-reciprocal Networks—H. J. Carlin. (PROC. IRE, vol. 43, pp. 608–616; May, 1955.) "The necessary and sufficient conditions are given that a matrix with arbitrary complex number elements be the impedance, admittance or scattering matrix of a physical linear reciprocal or non-reciprocal network. A canonical form for non-reciprocal network synthesis is presented which applies to any linear *n*-terminal pair (*n*-port) system at any fixed frequency. If the network is passive the only circuit element required in addition to lossless inductors, capacitors, transformers and positive resistors is the gyrator. If the network is active, negative resistors and gyrators must be used in addition to conventional passive elements. Some discussion of matrixes with frequency variable elements is also given."

621.372.5 2226
Transient Responses with Limited Overshoot—A. J. O. Cruickshank. (*Wireless Eng.*, vol. 32, pp. 154–163; June, 1955.) "The problem is considered of designing a system transfer function such that, in the response to a step-function input, the overshoot shall be limited and consist largely of the principal mode of oscillation. A method is stated by which the size of any real term in the response may be calculated relative to the maximum value of the principal mode. Graphical conditions are given restricting the permissible pole and zero positions in certain cases when this ratio is specified. The procedure is illustrated by examples."

621.372.5 2227
Design of Parallel-T Resistance-Capacitance Networks—Y. Oono. (PROC. IRE, vol. 43, pp. 617–619; May, 1955.) Formulas are derived for networks having a transfer characteristic symmetrical with respect to resonance frequency and a minimum loss for a prescribed frequency discrimintion.

621.372.5 2228
Symmetrical Quadripoles operated as [asymmetrical] Transformers—A. Ruhrmann. (*Arch. Elektrotech.*, vol. 41, pp. 320–333; November 25, 1954.) A theoretical investigation is presented of the behavior of passive, linear quadripoles in asymmetrical operation. The analysis is based on introduction of the characteristic iterative impedance. Multisection bandpass filters with different output and input impedances are built up with sections which have similar frequency characteristics and phase constants, but geometrically graduated impedances. A numerical example is given of a 30Ω/60Ω transformation for the 474–1,570-kc frequency band.

621.372.54 2229
The Initial Transient Behaviour of Filters with Characteristic Amplitude Response—H. H. Nissen and W. Händler. (*Arch. elekt. Übertragung*, vol. 9, pp. 74–80; February, 1955.) The distortion of a step-voltage input by Butterworth-type and Tchebycheff-type filters is analyzed. Calculations are made for all-pass equalizing networks for the former type.

621.372.54 2230
A Note on Time Series and the use of Jump Functions in Approximate Analysis—A. J. O. Cruickshank. (*Proc. IEE*, Part C, vol. 102, pp. 81–87; March, 1955. Digest, *ibid.*, Part II, vol. 101, p. 680; December, 1954.) The response of filters to jump-function inputs is investigated. Jump-transfer functions and serial operators for common types of filter are tabulated. The method is compared with that described by Tustin (*Jour. IEE*, Part IIA, vol. 94, p. 130; 1947.)

621.372.54+621.372.2]:512.831 2231
The Matrix Approach to Filters and Transmission Lines—M. E. Fisher. (*Electronic Eng.* vol. 27, pp. 198–204 and 258–263; May and June, 1955.) Use of matrices provides a

direct and flexible method of treatment. A simple graphical procedure is indicated for investigating filter characteristics.

621.372.54:519.241.1 2232

A New Method of Determining Correlation Functions of Stationary Time Series—Lampard. (*See* 2352.)

621.372.54:621.3.015.3 2233

The Transient Response of R.F. and I.F. Filters to a Wave Packet—A. W. Gent. (*Proc. IEE*, Part C, vol. 102, pp. 1–2; March, 1955) Discussion on 963 of 1953.

621.372.56:546.289:538.63 2234

A Novel Microwave Attenuator using Germanium—J. B. Gunn and C. A. Hogarth. (*Jour. Appl. Phys.*, vol. 26, pp. 353–354; March, 1955.) The device comprises a thin rectangular slab of Ge set obliquely across a waveguide, with current leads attached to the two ends and a magnetic field applied parallel to the face of the slab so as to control the concentration of current carriers as described by Weisshaar and Welker (3590 of 1954). Attenuation of 33 db was obtained with a pulsed control current of 60 ma.

621.372.6 2235

The Hybrid Ring—a Hybrid Circuit for Very High Frequencies—C. Colani. (*Frequenz*, vol. 8, pp. 368–372; December, 1954.) Input admittance and coupling attenuation are calculated as functions of frequency; good agreement with measured values is obtained. Methods of increasing bandwidth are discussed and a practical arrangement is described for the 4-kmc band, using a construction similar to microstrip.

621.373 2236

Suggestions on constructing Oscillators—J. Piesch. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.* vol. 9, pp. 10–19; January/February, 1955.) The operation of various tube and transistor oscillators is analyzed. Fixed and continuously tunable oscillators, pulse generators, and two-frequency oscillators for special applications are discussed.

621.373.2.029.65 2237

The Generation of Millimetre Waves—J. L. Farrands. (*Proc. IEE*, Part C, vol. 102, pp. 98–103; March, 1955. Digest, *ibid.*, Part B, vol. 102, p. 264; March, 1955.) Theory is presented for spark generators; experimental evidence generally supports the theory. Useful results can only be obtained with sparks in oil.

621.373.42.029.4 2238

Ultra-low-Frequency Oscillator—M. D. Armitage. (*Wireless Eng.* vol. 32, pp. 173–174; June, 1955.) A *CR* circuit using resistors and polystyrene capacitors of moderate values is described for obtaining oscillations at frequencies down to <1 cps. The effective value of the resistance is increased by use of the cathode-follower effect.

621.373.421 2239

Parallel-Network Oscillators—J. L. Stewart. (PROC. IRE, vol. 43, pp. 589–595; May, 1955.) General principles are discussed of electronically tunable wide-band oscillators having two signal paths, with their outputs added, and a common feedback path; the tuning is performed by differentially controlling the gain in the two paths, e.g. by push-pull modulation. Particular arrangements described include a twin-triode circuit using an artificial-transmission-line network, giving a 2:1 tuning ratio, and a four-tube circuit with lead and lag networks, giving a frequency range nearly independent of center frequency. Amplitude constancy, stability, waveform and noise are examined.

621.373.5:[621.314.7+621.372.412 2240

A Simple Quartz Crystal Oscillator driven by a Junction Transistor—H. G. Bassett. (*Electronic Eng.*, vol. 27, pp. 222–223; May, 1955.) The transistor is connected with grounded base and the quartz crystal is operated in its series mode. The frequency range is up to about 300 kc and the output power up to about 10 mw.

621.374.4 2241

Four-Decade Frequency Divider—G. K. Jensen and J. E. McGeogh. (*Electronics*, vol. 28, pp. 154–158; April, 1955.) Division of any frequency from sub-audio to 450 kc by any whole number up to 10,999 is accomplished by a direct-reading divider using binary counter circuits.

621.375.2:621.372.2 2242

Design of Coaxial Cavities for Valves with Planar Electrodes—L. Grifone. (*Alta Frequenza*, vol. 23, pp. 357–377; December, 1954.) The cavities form the double coaxial-line structure used in conjunction with a disk-seal-type tube for grounded-grid operation as amplifier or multiplier. A graphical method of design is presented, which involves only functions of the characteristic impedance, once the operating mode has been chosen. Criteria are explained for determining the diameters of the two cavities for optimum response. For higher-mode operation the response can be improved most easily by the introduction of a suitable discontinuity. Three types are considered. The occurrence of spurious resonances is also discussed.

621.375.2:621.396.41 2243

Design of H.F. and I.F. Amplifiers for Multichannel F. M. Links—R. Schienemann. (*Telefunken Ztg*, vol. 24, pp. 157–162 and 211–219; September and December, 1954. Correction, *ibid.*, vol. 24, p. 254; December, 1954.) Bandwidth required is determined (a) from the side-band amplitudes for given modulation index and (b) from the permissible distortion. Expressions for (b) involve the derivatives of the amplifier response curves and are tabulated for single-stage circuits and two-stage bandpass filters. The computation for multi-stage circuits is shown to involve the same derivatives. Under ideal conditions, second-harmonic distortion would be eliminated by exact tuning to the central frequency, and third-harmonic distortion sufficiency reduced by using symmetrical filters with a coupling coefficient $1/\sqrt{3}$. In practice, tuning is not exact, but second-harmonic distortion can be kept at a tolerable level by adjustment of the pass-band response at the alignment stage. Tube capacitance variations are allowed for in the shunt circuit capacitance. Neutralization can practically eliminate the effects of feedback via grid-anode capacitance. Criteria for the choice of IF are explained.

621.375.2.029.3 2244

Circuit Design Factors for Audio Amplifiers—M. V. Kiebert, Jr. (*Electronics*, vol. 28, pp. 166–171; April, 1955.) Close regulation of the anode supply of a cathode-follower driver stage in a power amplifier is achieved by using a regulator triode operated by a diode connected to the driven grid. Circuits discussed include improved versions of Williamson and "ultra-linear" amplifiers.

621.375.2.029.3 2245

Design for a 20-Watt High-Quality Amplifier: Part 2—W. A. Ferguson. (*Wireless World*, vol. 61, pp. 279–282; June, 1955.) Details are given of the circuit, lay-out and performance of an amplifier with an output stage of two Type-EL34 high-slope pentodes in push-pull with partial screen-grid loading. Part 1: 1918 of August.

621.375.227 2246

A New Circuit for balancing the Characteristics of Pairs of Valves—R. E. Aitchison. (*Electronic Eng.*, vol. 27, pp. 224–226; May, 1955.) A more detailed account of the work described previously (982 of May).

621.375.23.029.64 2247

X-Band Receiving Amplifier—K. Ishii. (*Electronics*, vol. 28, pp. 202, 210; April, 1955.) A regenerative klystron amplifier is described, suitable for use in a microwave television receiver, having a gain 16 db at 9.76 kmc, with a bandwidth of 20 mc.

621.375.3 2248

Problems of Magnetic Pre-amplifiers—F. Kümmel. (*Elektrotech. Z.*, *Edn. A*, vol. 76, pp. 113–120; February 1, 1955.) The topics discussed include the quality, the effect of core shape on the magnetization characteristic, effect of characteristic of tube or semiconductor rectifier on amplification, and effects of incomplete magnetic decoupling of input circuit.

GENERAL PHYSICS

53(083.7) 2249

Representation of Physical Quantities in Formulae, Tables and Coordinate Systems—C. Glinz. (*Tech. Mitt. schweiz. Telegr. Teleph-Verw.*, vol. 33, pp. 41–69; February 1, 1955. In German.) A comprehensive discussion covering both fundamental physical principles and typographical aspects.

53.081.4 2250

A New System of Logarithmic Units—J. B. Moore. (PROC. IRE, vol. 43, p. 622; May, 1955.) Comment on 988 of May (Hartley).

530.112 2251

Proposal for a New Aether Drift Experiment—D. D. Crombie. [*Nature* (*London*), vol. 175, pp. 350–351; February 19, 1955.] A method of greater sensitivity than Essen's (2618 of 1954) uses a resonant cavity to produce a phase change in a signal from a fixed-frequency oscillator.

530.145:538.3 2252

Derivation of the Laws of Relativistic Electrodynamics for a Vacuum from the Energy-Quantum Model—H. Zuhrt. (*Arch. elekt. Übertragung*, vol. 9, pp. 47–51; January, 1955.) Theory developed previously (1627 of July) is extended to include the case of systems in motion.

530.145.6:621.385.833 2253

Wave-Mechanics Theory of Electron-Optical Image Formation: Part I—W. Glaser and G. Braun. (*Acta Phys. austriaca*, vol. 9, pp. 41–74; December, 1954.)

537.21:621.3.013.78 2254

Electrostatic Averaged Boundary Conditions for Wire Mesh—B. Ya. Moyzhes. (*Zh. Tekh. Fiz.*, vol. 25, pp. 167–176; January, 1955.) A brief review is given of the existing methods for determining the screening effect of a mesh, and a general derivation is proposed of the boundary condition for a mesh by the method of averaging the field. The results obtained are applied to the determination of the screening effect of an earthed uniform spherical mesh with respect to a charge or a dipole placed at the center, and also to the determination of the effective potential in a triode. See also 2273 below.

537.212:621.317.39.082.72 2255

The Field due to an Infinite Dielectric Cylinder between two Parallel Conducting Planes—C. Mack. (*Brit. Jour. Appl. Phys.*, vol. 6, pp. 59–62; February, 1955.) If the cylinder radius, b, is sufficiently small compared with the other dimensions involved, the change of capacitance due to its insertion between the plates can be made proportional to $(K-1)b^2/(K+1)$, where K is the dielectric constant of the cylinder. For a conducting cylinder, K is put equal to infinity. Applications include the measurement of the departure from uniformity of threads.

537.226:537.311.35 2256
The Ionic Conductivity of Dilute Potassium Chloride Solutions at Centimetric Wavelengths—V. I. Little and V. Smith. (*Proc. Phys. Soc.*, vol. 68, pp. 65—74; February 1, 1955.) Measurements on solutions in the concentration range 0.005–1.0 normal, at a frequency of 3×10^9 cps, indicate the existence of a strong dispersion region at concentrations below 0.5 normal, which may be explained in terms of the perturbations by the applied field of a shell of water molecules surrounding the ion at a mean distance of 6 Å.

537.226.2:546.212:621.317.335.3 2257
The Dielectric Constant of 'Free' and 'Bound' Water at Microwave Frequencies—J. Baruch and W. Low. (*Bull. Res. Coun. Israel*, vol. 3, pp. 31–36; June–September, 1953.) An attempt is made to assess and summarize results obtained by the authors and by other workers on the dielectric dispersion of water, the water molecules in $CuSO_4\cdot5H_2O$ in the temperature range from −70 degrees C. to 150 degrees C., water in hygroscopic salts, aqueous gels and thixotropic gels. Estimation of hydration is also considered. 39 references.

537.311.1:535.137 2258
The Theory of the Reflectivity of Metals—R. Wolfe. (*Proc. Phys. Soc.*, vol. 68, pp. 121–127; February 1, 1955.) The reflectivity of an ideal metal for infrared radiation is calculated by quantum-mechanicals. Results agree with those obtained by Dingle (2626 of 1954) using classical methods; the diffuse-reflection wave functions correspond more closely than the specular-reflection wave functions to the actual electronic wave functions as indicated by experimental results.

537.311.6:621.315.514 2259
Skin and Spiraling Effect in Stranded Conductors—J. Zaborszky. (*Trans. Amer. IEE*, Part III, *Power Apparatus and Systems*, vol. 72, pp. 599–602; August, 1953. Discussion, pp. 602–603.) An analysis is made to determine the ac resistance and internal reactance of stranded conductors with or without a magnetic core, at power frequencies, assuming current density constant in individual strands. The spiralling of strands tends to reduce skin effect. Data calculated for standard copper conductors with parallel and spiralled strands are listed.

537.52:537.226 2260
Theory of Breakdown of Inhomogeneous Dielectrics—Yu. M. Volokobinski. (*Zh. Tekh. Fiz.*, vol. 25, pp. 74–80; January, 1955.)

537.525 2261
Formative Time-Lag Studies with High-Frequency Discharges—A. W. Bright and H. C. Huang. (*Proc. IEE*, Part C, vol. 102, pp. 42–45; March, 1955. Digest, *ibid.*, Part III, vol. 101, pp. 407–408; November, 1954.) Pulse techniques were used to study phenomena in coaxial-cylinder and parallel-wire systems, in the frequency range 1–15 mc, where the discharge changes from corona type to one which completely crosses the gap. The breakdown voltage drops by about 13 per cent above the critical frequency.

537.525 2262
Space-Charge Effects in a High-Frequency Discharge: Part 2—M. Chenot. (*Jour. Phys. Radium*, vol. 16, pp. 101–107; February, 1955.) Further discussion of results of experiments previously reported (1945 of August). Cathode disintegration phenomena, various shapes of characteristic, and factors affecting the direction and magnitude of the emf are considered.

537.533 2263
Field-Emission and Surface Phenomena—F. L. Jones. [*Nature* (*London*), vol. 175, pp. 244–245; February 5, 1955.] Brief report of a symposium held at Pittsburg in November, 1954.

537.533 2264
Delayed Electron Emission from Metals—K. Seeger. (*Naturwiss.*, vol. 42, p. 66; February, 1955.) Measurements were made on a w foil substituted for the usual photocathode in a multiplier cell. Emission was excited by bombarding with 1-kev electrons and measurements were started 5 minutes after the bombardment, the foil being simultaneously heated at a controlled rate. The emission/temperature curve passed through a maximum between 130 degrees and 200 degrees C. Emission was eliminated by prolonged heating at 1,000 degrees C. but was restored by exposing the foil to the atmosphere at 100 degrees C. and abrading it. The results support theory advanced by Nassenstein (87 of February).

537.533 2265
Electron Emission from Metal Surfaces after Mechanical Working—J. Lohff and H. Raether. (*Naturwiss.*, vol. 42, pp. 66–67; February, 1955.) Measurements were made on various metals in gas at a pressure of about 2×10^{-5} Torr, using an electron multiplier. The surfaces were abraded with a steel brush within the vessel. Emission/time curves are shown for Pb, Ca, Al and Na; the activity of the elements is correlated with their position in the periodic system. Oxidation may be the cause of the emission, but no explanation of the mechanism is yet available.

537.533.8 2266
On the Escape Mechanism of Secondary Electrons from Insulators—A. J. Dekker. (*Physica*, vol. 21, pp. 29–38; January, 1955.) Starting from the Boltzmann transport equation, and with certain simplifying assumptions, theory is developed which takes into account the interaction of secondary electrons with lattice vibrations, traps and occupied donor levels. At high trap densities, the secondary yield for high primary energies is inversely proportional to the square root of the trap density, at low trap densities yield is independent of trap density. Expressions for the temperature effect for high primary energies are derived for polar and nonpolar insulators. Results for the first case have been confirmed by experiment.

538.11 2267
Antiferromagnetism—G. W. Pratt, Jr. (*Phys. Rev.*, vol. 97, pp. 926–932; February 15, 1955.) The nature of the spin coupling in the antiferromagnetic oxide MnO is discussed on the basis of a simplified model. The coupling between magnetic ions whose charge densities interact with an intervening nonmagnetic ion but not directly with each other is described as the result of the polarization of the nonmagnetic ion.

538.56 2268
Representation of Electromagnetic Field by Retarded Potentials—I. S. Arzhanykh. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 100, pp. 1053–1056; February 21, 1955. In Russian.] Applications considered include a boundary problem and the equation of motion of electron gas in vacuo.

538.566+534.2 2269
Electromagnetic and Acoustic Scattering by a Semi-infinite Body of Revolution—C. E. Schensted. (*Jour. Appl. Phys.*, vol. 26, pp. 306–308; March, 1955.) Theory developed by Kline (646 of 1952) is used to determine the scattering of a plane wave incident along the axis of a perfectly reflecting body. For the case of a paraboloid the solution of the em problem is obtained exactly in closed form. The results are compared with those obtained from consideration of current distribution, as in 2306 below.

538.566+534.2 2270
Electromagnetic and Acoustical Scattering from a Semi-infinite Cone—K. M. Siegel, J. W. Crispin and C. E. Schensted. (*Jour. Appl. Phys.*, vol. 26, pp. 309–313; March, 1955.) Calculations of scattering cross section for em waves incident along the axis of the cone are made by the physical-optics method based on current distribution and by the exact method based on the appropriate field equations; the solutions are practically identical for cone semi-angles close to zero or $\pi/2$. The corresponding comparison is also made for acoustic waves.

538.566:535.42 2271
Studies of the Diffraction of Electromagnetic Waves by Circular Apertures and Complementary Obstacles: the Near-Zone Field—M. J. Ehrlich, S. Silver and G. Held. (*Jour. Appl. Phys.*, vol. 26, pp. 336–345; March, 1955.) Measurements have been made of both magnetic and electric field strengths with diffracting apertures of diameter up to 36λ and normally incident waves. Experimental techniques for illuminating the aperture and probing the field are described; wavelengths around 3 cm afford a satisfactory compromise between reasonable over-all dimensions on the one hand and reliability of measurements on the other. Results are shown graphically; the predicted uniformity of the tangential magnetic field in the aperture plane is confirmed.

538.566:537.5:523.72 2272
Nonlinear Theory of Space-Charge Wave in Moving, Interacting Electron Beams with Application to Solar Radio Noise—H. K. Sen. (*Phys. Rev.*, vol. 97, pp. 849–855; February 15, 1955.) Consideration of the exact equations for interaction between two neutralized electron beams indicates that steady-state space-charge waves can be propagated in such a medium. The period of the wave depends on the amplitude and phase velocity. The oscillation is simple harmonic for small amplitudes but deviates progressively from this condition as the amplitude increases. Beyond a critical amplitude, whose value depends on the phase velocity, the oscillation waveform becomes discontinuous. The theory is used to estimate the relative intensity of the second-harmonic component in solar rf outbursts reported by Wild *et al.* (391 of 1954). Analysis based on theory of radiation from oscillating plasma leads to a value for the rf flux of the order of magnitude of that observed.

538.566:621.3.013.78 2273
Electrodynamic Averaged Boundary Conditions for Wire Mesh—B. Ya Moyzhes. (*Zh. Tekh. Fiz.*, vol. 25, pp. 158–166; January, 1955.) Eqs. (29) and (30) are derived which form a system of boundary conditions relating the averaged tangential components of the electric and magnetic fields on the surface of the mesh to the averaged surface density of the current in the mesh. The results obtained are illustrated by calculating the reflection of a plane wave from a plane uniform mesh, and by determining the screening effect of a spherical mesh with respect to a small turn of wire carrying a current and located at the center of the sphere.

538.566.2:535.32:[546.212+546.332.26 2274
The Refractive Index of Water and Aqueous Solutions of Sodium Sulphate at Metre Wavelengths—I. S. Frolov. (*Zh. Eksp. Teor. Fiz.*, vol. 27, pp. 477–486; October, 1954.) Using the first method of Drude, the determination of the dependence of the refractive index of water on wavelength and temperature was extended to meter wavelengths and a temperature range from 2 degrees to 32 degrees C. Similar investigations were also carried out to determine the dependence on concentration of the refractive index of aqueous solutions of Na_2SO_4. The main conclusions reached are: (a) the refractive index in both cases increases linearly with temperature; (b) there is no anomalous dispersion at meter wavelengths; (c) for a given temperature the refractive index of a solution increases with concentration.

538.569.4.029.64:535.338 2275
Application of Molecular Beams to Radio-spectroscopic Study of Rotation Spectra of Molecules—N. G. Basov and A. M. Prokhorov. (*Zh. Eksp. Teor. Fiz.*, vol. 27, pp. 431–438; October, 1954.) Molecular beams may be used to obtain narrow spectral lines, of width about 7 kc, and the rotation spectra of materials in the solid state. Quantitative estimates are made of the possibility of detecting the rotation transition $J=1 \rightarrow J=2$ in CsF molecules at a frequency of 17.7 kmc using a spectroscope with a waveguide absorption cell, and of the $J=0 \rightarrow J=1$ transition using a cavity-resonator instrument [see also 100 of February (Gordon *et al.*)].

539.233/.234 2276
Growth of Thin Films—G. Cario and J. H. Kallweit. (*Z. Phys.*, vol. 140, pp. 47–56; January 13, 1955.) An attempt is made to explain qualitatively the mechanism of growth on the basis of Frenkel's theory of adsorption. A summary is given of the most important parameters which may affect the structure of the film before, during, and after deposition.

548.0:53 2277
Investigation of the Natural Frequencies and Natural Oscillations of Imperfect Crystal Lattices—E. Fues and H. Stumpf. (*Z. Naturf.*, vol. 9a, pp. 897–902; October, 1954.) A theoretical investigation.

538.3 2278
Electromagnetics [Book Review]—J. D. Kraus. Publishers: McGraw-Hill, London, 1953, 604 pp., 76s.6d. [*Nature* (*London*), vol. 175, pp. 358–359; February 26, 1955.] A treatment of Maxwell's theory and some of its applications; suitable as a reference book for research workers. Mathematical functions entering into the solution of wave equations are dealt with in an appendix.

538.569.4:535.33 2279
Microwave Spectroscopy [Book Review]—W. Gordy, W. V. Smith and R. F. Trambarulo. Publishers: J. Wiley & Sons, New York, and Chapman & Hall, London, 1953, 446 pp., 64s. [*Nature* (*London*), vol. 175, p. 273; February 12, 1955.] Includes chapters on spectroscope technique and design, on measurements on gases, liquids and solids, on nuclear properties, on molecular structure, and on further possible applications of microwaves.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 2280
The Diffusion of Ionized Meteor Trails in the Upper Atmosphere—J. S. Greenhow and E. L. Neufeld. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 133–140; March, 1955.) The variation with height of the duration of radio echoes from meteor trails is discussed. For the experimental conditions involved, echo duration decreases approximately exponentially with increased height, in agreement with diffusion theory. The diffusion coefficient is between 10^4 and 10^5 cm^2, in agreement with estimates based on rocket determinations of upper-air density.

523.5:537.564 2281
Collision Processes in Meteor Trails—H. S. W. Massey and D. W. Sida. (*Phil. Mag.*, vol. 46, pp. 190–198; February, 1955.) Momentum-loss cross sections for impacts between meteor atoms and atmospheric molecules are calculated. The values of the ionization cross section derived from them are in agreement with those derived earlier by Bates and Massey (*Phil. Mag.*, vol. 45, pp. 111–122; February, 1954.) The diffusion of ionization in a meteor trail is also discussed in relation to known mobilities of positive ions in gases.

523.72:538.566:537.5 2282
Nonlinear Theory of Space-Charge Wave in Moving, Interacting Electron Beams with Application to Solar Radio Noise—Sen. (See 2272.)

523.72:621.396.822.029.62 2283
Character of 200-Mc/s Solar Noise Observation Equipment installed at Hiraiso Radio Wave Observatory—T. Takahashi, M. Onoue and K. Kawakami. [*Jour. Radio Res. Labs.* (*Japan*), vol. 1, pp. 41–53; September, 1954.] Principles of measurement and accuracy of equipment installed in March, 1952 are discussed. Routine observations started in August, 1954. The path of the sun is followed by a beam antenna system rotated 2 degrees every eight minutes, with manual adjustment daily for change of declination. During one minute in every eight the receiver input is connected to a noise generator for calibration purposes. The double-superheterodyne receiver has a bandwidth of about 80 kc. Results are compared with data from other stations.

523.74+551.510.535]:537.311.37 2284
Solar Electrodynamics—J. W. Dungey. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 88–90; March, 1955.) The reduction in conductivity of an ionized gas by a magnetic field and the effect of a polarization electric field are discussed. The theory has a bearing on conditions in the solar atmosphere above sunspots and on tidal oscillations in the ionosphere.

523.75:[551.510.535+550.38 2285
Geophysical Aspects of Solar Flares—V. C. A. Ferraro. [*Nature* (*London*), vol. 175, pp. 242–244; February 5, 1955.] Report of a discussion held at the Royal Astronomical Society in November, 1954. Results of research on the geomagnetic and ionospheric effects of the flares were surveyed.

523.755:523.16:621.396.822:523.99 2286
The Irregular Structure of the Outer Regions of the Solar Corona—A. Hewish. (*Proc. Roy. Soc. A*, vol. 228, pp. 238–251; February 22, 1955.) Records obtained in 1952 of the radiation from the radio star in Taurus [98 of 1953 (Machin and Smith)] and results of measurements made in June, 1953, using antennas of greater directivity, cannot be explained in terms of absorption or large-scale refraction effects in the solar corona, but are consistent with a scattering theory. An estimate is made of the size and the electron density of the coronal irregularities in the range of distance 5–15 $R_\odot$, where $R_\odot$ is the solar radius. The irregular structure may represent an extension of the visible coronal rays.

550.384 2287
Recent Progress in the Theory of the Main Geomagnetic Field—S. K. Runcorn. (*Sci. Prog.*, vol. 43, pp. 13–27; January, 1955.)

550.386:523.75 2288
Particularly Large Chromospheric Eruptions and Geomagnetic Activity—P. Simon. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1056–1058; March 7, 1955.] Further analysis of observations indicates that geomagnetic activity related to the appearance of certain eruptions can be distinguished from that related to the central meridional passage of certain sunspots (1973 of August).

551.510.5 2289
Ion Production Rate in an Atmosphere of Exponentially Rising Temperature with Height—H. Kamiyama. (*Sci. Rep. Tohoku Univ.*, *5th Ser.*, *Geophys.*, vol. 6, pp. 11–18; August, 1954.) An expression is derived for the height distribution of the ion production rate, assuming that temperature varies exponentially with height and also depends on the solar zenith distance. Results are similar to those of Chapman for heights up to about 100 km. In the F_2 region the maximum rate of ion production and the height at which it occurs vary anomalously with solar zenith distance.

551.510.535 2290
Mechanism of Creation of Ionospheric Inhomogeneities—B. N. Gershman and V. L. Ginzburg. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 100, pp. 647–650; February 1, 1955. In Russian.] Results of approximate calculations indicate that a negative temperature gradient of about 10^{-4} degrees per cm, or greater, at a height of about 400 km could result in thermal convection currents such as would produce inhomogeneities and high-velocity winds in the ionosphere. The temperature at 300–400 km is assumed to be about 1,000 degrees–3,000 degrees.

551.510.535 2291
Studies on the Sunrise Effect in Regions E and F—S. S. Baral. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 160–170; March, 1955.) To be effective in causing ionization in a night-time layer, solar radiation must pass above the day-time height of the layer. The observed delay in time of sunrise effect is due therefore to the difference between the night-time and day-time heights of the layer considered. Observations made at Calcutta over the period 1947–1952 are in reasonable agreement with the theory.

551.510.535 2292
Absence of Bifurcation in the E Layer—J. C. Seddon and J. E. Jackson. (*Phys. Rev.*, vol. 97, pp. 1182–1183; February 15, 1955.) Results of rocket measurements of electron concentration [see e.g. 1033 of May (Seddon et al.)] are adduced as evidence of absence of the bifurcation reported by Lien et al. (1054 of 1954).

551.510.535 2293
Further Remarks on Bifurcation in the E Layer—W. Pfister, J. C. Ulwick and R. J. Marcou. (*Phys. Rev.*, vol. 97, pp. 1183–1184; February 15, 1955.) Comment on 2292 above.

551.510.535 2294
Horizontal Movements of Ionization in the Equatorial F Region—B. W. Osborne. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 117–123; March, 1955.) Results of measurements at Singapore during the period September, 1953–April, 1954 show the existence of a regular semidiurnal sinusoidal variation in the *E-W* velocity component at the equinoxes. Around the December solstice, during the mornings, the movements were irregular. No correlation with virtual height was found, nor any regular N-S movement.

551.510.535 2295
Origin of the F_1 Layer—E. Chvojková. (*Bull. Astr. Inst. Csl.*, vol. 4, pp. 101–109; September 1, 1953. In German.) The splitting of the *F* layer into the F_1 and F_2 layers is considered to be due to a thermal effect. The maximum temperature occurs in the region of maximum ion production, but since the density of gas in this region is a minimum, higher electron concentrations will occur above and below it. The region of maximum ionization lies in the lower half of the *F* layer and hence the F_1 layer will be thinner than the F_2 layer. The formula derived for the height distribution of the electron concentration shows that the thermal process could lead to splitting of an ionized layer and also explains the secondary nocturnal maximum of $f^{\circ}F$. The peculiarities of the equatorial layer described by Osborne (989 of 1952) are briefly discussed.

551.510.535:001.4 2296
Nomenclature and Conventions used in Analysis of Ionospheric Data—K. Bibl, R. Busch, K. Rawer and K. Suchy. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 69–87; March, 1955, In French.) It is proposed to classify the layers according to their thickness. A "critical thickness" is introduced to define the difference between normal layers, which are always thick, and abnormal thin layers, caused by transitory phenomena. The term "critical frequency" may

be used of thin layers if account is taken of the intensity of the echoes. Characteristics directly determined and those found by extrapolation should be distinguished. Revised lists of characteristic symbols and code letters are given in appendices.

551.510.535:523.746 2297

Relations between the Critical Frequency of the Ionosphere F_2 Layer at Freiburg and Solar Activity Centres during the Years 1948–1951—P. Simon. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1192–1193; March 14, 1955.] Analysis is made separately for night-time and daytime observations. No significant variation of F_2-layer critical frequency was associated with the passage of those sunspots giving no rf radiation. On passage of sunspots giving rf radiation, the night-time critical frequency suffered a significant reduction, while the daytime value was not significantly affected. This night-time effect is in concordance with observations of ionospheric absorption reported by Davies and Hagg (1765 of July).

551.510.535:523.78 2298

Ionospheric Behaviour at Khartoum during the Eclipse of 25th February 1952—C. M. Minnis. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 91–112; March, 1955.) Observations indicate that the E and F_1 layers behave as Chapman layers having constant effective recombination coefficients α' of 1.5×10^{-8} cm³ s⁻¹ and 8×10^{-9} cm³ s⁻¹ respectively, assuming a nonuniform distribution of the sources of ionizing radiation on the sun's disk. The positions and relative intensities of these sources and the frequency of the radiation are indicated. The F_2-layer records suggest the existence at the time of the eclipse of two component layers: a lower one corresponding to the normal post-sunrise F_2 layer, having a high α', and an upper one which is a later development of the lower, having a low α'. Changes in D-later absorption confirm the asymmetrical distribution of sources of ionizing radiation. There is no evidence of an eclipse effect on E_s ionization, nor for a corpuscular eclipse in the F_2 layer.

551.510.535:621.317.39 2299

An Apparatus for recording Time Delays between Radio Fading Characteristics—G. J. Phillips. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 124–128; March, 1955.) The equipment is designed for the study of ionospheric winds (1916 of 1952) and records time delays in the range 0.1 to 5s, with an instrumental accuracy of about 0.1s.

551.510.535:621.396.11.029.62 2300

Sporadic—E Propagation—Gerson. (*See* 2409.)

551.510.535:621.396.6:621.317.7 2301

D.S.I.R. Ionospheric Absorption Measuring Equipment—W. R. Piggott. (*Wireless Eng.*, vol. 32, pp. 164–169; June, 1955.) Main design features are described of pulse transmitting and receiving equipment used at Slough since 1941. Highly stable operation is essential. Pulse duration is variable between 50 μs and 500 μs; the frequency range is 1–10 mc, and a peak power output of 0.5 kw is adequate for observer reception. Rhombic, delta and folded-dipole transmitting antennas have been used. The receiver is a superheterodyne with extra-high gain and dynamic range, and with provision for attenuation varying continuously up to about 160 db. Double-doublet and folded-dipole receiving antennas have been used. The display unit is a conventional cro.

551.593 2302

Collisional Deactivation and the Night Airglow—D. R. Bates. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 171–172; March, 1955.) A possible reaction of the oxygen is indicated to explain the observed spectrum of the night airglow.

551.510.53 2303

The Earth's Exterior Atmosphere and the Counterglow: The Counterglow as Related to Modern Geophysical Theories, with Seven Recent Russian Papers. Collected and translated by E. R. Hope [Book Review]—Publishers: Defence Research Board, Ottawa, 2nd ed., 1954, 52 pp. [*Nature* (*London*), vol. 175, p. 377; February 26, 1955.] Theory based on observations suggests that a gaseous tail resembling that of a comet stretches away from the earth in the direction away from the sun. A valuable contribution to the literature of upper-atmosphere research.

LOCATION AND AIDS TO NAVIGATION

621.396.663:621.396.933.1 2304

Qualitative and Quantitative Errors of an Automatic Radio Compass—A. Troost and G. Ziehm. (*Fernmeldetech. Z.*, vol. 8, pp. 65–70; February, 1955.) Three sources of bearing errors in a typical automatic radio compass are considered: (a) effect of "nondirectional" noise voltage in goniometer output, (b) effect of noise voltage and an error phase angle between the goniometer output voltage and the output voltage of the omnidirectional aerial, and (c) effect of inaccurate tuning.

621.396.933:551.594.6 2305

The Measurement of Low-Frequency Atmospheric Noise in Southern Africa—(*ICAO Bull.*, vol. 9, pp. 7–8; October, 1954.) General account of work undertaken since 1946 for the development of radio navigational aids in South Africa. This includes the recording of atmospheric noise in aircraft and on the ground at 100 kc [438 of 1951 (Hogg)], and ground tests carried out by the Telecommunications Research Laboratory on Decca, radio-compass, consol and loran equipment installed at fixed sites; in these, atmospheric noise picked up by an antenna and measured on a recorder was fed to the equipment together with a signal to simulate actual operating conditions.

621.396.96:538.566 2306

Bistatic Radar Cross-Sections of Surfaces of Revolution—K. M. Siegel, H. A. Alperin, R. R. Bonkowski, J. W. Crispin, A. L. Maffett, C. E. Schensted and I. V. Schensted. (*Jour. Appl. Phys.*, vol. 26, pp. 297–305; March, 1955.) The "bistatic" cross section $\sigma(\beta)$ corresponds to an arrangement in which radar transmitter and receiver are separated, β being the angle subtended at the scatterer by the transmitter-receiver line. Simple geometrical configurations are considered in which the transmitter is located on the axis of symmetry of the scatterer. Formulas are derived, by considering current distribution, for scatterers of spheroidal, conical, ogive, paraboloidal, ellipsoidal and hyperboloidal form, with characteristic dimensions large compared with λ. *See* also 2269 above.

621.396.96.089.6 2307

Precision Calibrator checks Radar Beacons—R. D. Sinish. (*Electronics*, vol. 28, pp. 150–153; April, 1955.) Description of portable transponder equipment which checks range accuracy within ±5 yards.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 2308

A Compensated Ionization Manometer for Measurement of Vacua with High Relative Accuracy—J. W. Hiby and M. Pahl. (*Z. Naturf.*, vol. 9a, pp. 906–907; October, 1954.)

535.37 2309

Action of Ni and of Co on ZnS(Au) Phosphors. Displacement of the Fermi Level by Variation of the Concentration—N. Arpiarian. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1333–1335; March 21, 1955.]

537.226:546.212 2310

The Dielectric Method of Studying the Adsorption of Water—J. Le Bot and S. Le Montagner. (*Jour. Phys. Radium*, vol. 16, pp. 163–164; February, 1955.) Continuation of experiments reported in 1998 of August using various types of gel and varying water content. Comparable results were obtained.

537.226.2 2311

Dielectric Mixture Chart—E. Sion. (*Electronics*, vol. 28, p. 176; April, 1955.) Nomogram relating dielectric constants of mixtures of two dielectrics with the dielectric constants of the components and the composition of the mixture.

537.227 2312

X-Ray and Neutron Diffraction Study of Ferroelectric $PbTiO_3$—G. Shirane, R. Pepinsky and B. C. Frazer. (*Phys. Rev.*, vol. 97, pp. 1179–1180; February 15, 1955.)

537.311.3+536.21 2313

Thermal and Electrical Conductivities of Solids at Low Temperatures—G. K. White and S. B. Woods. (*Canad. Jour. Phys.*, vol. 33, pp. 58–73; February, 1955.) Measurements were made over the range 2 degrees–300 degrees K on materials for which lattice conductivity is comparable to electronic conductivity. Results are presented for some dilute Cu alloys, Be, Bi and Ge.

537.311.3:546.3-1 2314

Semiconducting Properties of some High-Resistance Metallic Alloys—A. I. Drabkin. (*Zh. Tekh. Fiz.*, vol. 25, pp. 81–84; January, 1955.) A table is given showing the properties of various types of manganin, constantan and NWB alloys (Ag and Mn with Sb or Ni).

537.311.33+621.314.7 2315

Semiconductors and the Transistor—E. W. Herold. (*Jour. Frank. Inst.*, vol. 259, pp. 87–106; February, 1955.) A general survey with 33 references.

537.311.33 2316

Motion of Electrons and Holes in Perturbed Periodic Fields—J. M. Luttinger and W. Kohn. (*Phys. Rev.*, vol, 97, pp. 869–883; February 15, 1955.) "A new method of developing an effective-mass" equation for electrons moving in a perturbed periodic structure is discussed. This method is particularly adapted to such problems as arise in connection with impurity states and cyclotron resonance in semiconductors such as Si and Ge. The resulting theory generalizes the usual effective-mass treatment to the case where a band minimum is not at the center of the Brillouin zone, and also to the case where the band is degenerate. The latter is particularly striking, the usual Wannier equation being replaced by a set of coupled differential equations."

537.311.33 2317

Scattering of Electrons by Charged Dislocations in Semiconductors—W. T. Read, Jr. (*Phil. Mag.*, vol. 46, pp. 111–131; February, 1955.) Experiments [2124 of 1954 (Pearson et al.)] suggest that dislocations in Ge act as rows of closely spaced acceptor centers. In n-type material a dislocation accepts electrons and becomes a line of negative charge surrounded by a cylindrical region of fixed positive space charge. A theoretical investigation is made of drift and Hall mobilities for current flow normal to the dislocations, with a magnetic field (a) parallel to the dislocations, and (b) normal to them.

537.311.33 2318

Dielectric Properties of Transition Layers in Semiconductors—B. M. Vul. (*Zh. Tekh. Fiz.*, vol. 25, pp. 3–10; January, 1955.) Formulas are derived for determining the capacitance, resistance and dielectric losses of the region in which the semiconductor changes from n to p type.

537.311.33:537.32 2319
On the Thermoelectric Properties of the Impurity Semiconductors—E. Laurila. (*Ann. Acad. Sci. Fenn.*, Ser. A, no. 178, 11 pp.; 1955.) Formulas for the Seebeck, Peltier and Thomson coefficients of impurity semiconductors are derived on the basis of statistical theory of conductivity as developed by Shockley (246 of 1951). For a Ge *p-n* junction at room temperature the thermoelectric power is found to be about 2 mv/degrees K.

537.311.33:546.28 2320
Hyperfine Splitting of Donor States in Silicon—W. Kohn and J. M. Luttinger. (*Phys. Rev.*, vol. 97, pp. 883–888; February 15, 1955.) Calculations are reported which support the view that spin resonances observed by Fletcher et al. (3254 of 1954 and 453 of February) are due to electrons in donor states.

537.311.33:546.28:548.0 2321
Lamellar Defects in Single Crystals of Silicon—J. Franks, G. A. Geach and A. T. Churchman. (*Proc. Phys. Soc.*, vol. 68, pp. 111–112; February 1, 1955.) "Otherwise perfect single crystals of silicon have been shown to contain lamellar defects lying on (111) and (123) planes. Similar lamellae may be introduced by plastic deformation."

537.311.33:546.289 2322
Effective Carrier Mobility in Surface Space-Charge Layers—J. R. Schrieffer. (*Phys. Rev.*, vol. 97, pp. 641–646; February 1, 1955.) "Carriers held to a region near the surface by the potential well of a space charge layer may have their mobility reduced by surface scattering, if the width of the well is of the order of a mean free path. An effective mobility, which may differ from the bulk mobility by as much as a factor of ten, has been obtained from a solution of the Boltzmann equation. Solutions have been carried out for two types of potential functions: (*a*) a linear potential corresponding to a constant space-charge field, and (*b*) a solution of Poisson's equation including an external bias applied normal to the surface. The results have been used to calculate changes in surface conductance of germanium with changes in surface potential and predict the 'field effect' and 'channel effect' mobilities."

537.311.33:546.289 2323
Gold as an Acceptor in Germanium—W. C. Dunlap, Jr. (*Phys. Rev.*, vol. 97, pp. 614–629; February 1, 1955.) Measurements are reported on wafers cut from single crystals grown from a Ge melt to which Au had been added. Acceptor levels are found at 0.15 ev above the valence band and at 0.2 ev below the conduction band. At 77 degrees K, specimens of *p*- or *n*-type, with either high or low resistivity can be obtained, depending on the amounts of other impurities present. Photoconductive response extends to a wavelength of about 8 μ. Trapping phenomena are discussed. Important applications of Au-doped Ge in research are indicated.

537.311.33:546.289 2324
Properties of Germanium Doped with Cobalt—W. W. Tyler, R. Newman and H. H. Woodbury. (*Phys. Rev.*, vol. 97, pp. 669–672; February 1, 1955.) "Measurements of the temperature dependence of electrical resistivity in *n*- and *p*-type cobalt-doped germanium crystals indicate that cobalt introduces acceptor levels in germanium at 0.31 ± 0.01 eV from the conduction band and 0.25 ± 0.01 eV from the valence band. Ionization energies deduced from infrared photoconductivity studies at 77° K are in good agreement with the values obtained from resistivity measurements. *N*-type samples show higher intrinsic photosensitivity than *p*-type samples and demonstrate quenching effects."

537.311.33:546.289 2325
Sparked Hydrogen Treatment of Germanium Surfaces—N. Holonyak, Jr., and H. Letaw, Jr. (*Jour. Appl. Phys.*, vol. 26, p. 355; March, 1955.) On subjecting Ge specimens to a discharge in hydrogen at low pressure, the surface recombination velocity was greatly increased while the magnitude and decay time of voltages produced by illuminating *p-n* junctions was reduced; hole-storage effects in junctions were also reduced, without greatly impairing the reverse I/V characteristics.

537.311.33:546.47.86 2326
Diffusion of Sb and Sn in Semiconducting Compound SbZn—B. I Boltaks. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 100, pp. 901–903; February 11, 1955. In Russian.] Report of an experimental investigation. A sharp change of slope of the log-$D/(1/T)$ curve was observed at about 400 degrees C. This effect is briefly discussed.

537.311.33:546.472.21 2327
Investigations on Zinc Sulphide Crystals—J. Krumbiegel. (*Z. Naturf.*, vol. 9a, pp. 903–904; October, 1954.) Brief report of an investigation of the structure, luminescence and photoconductivity of single crystals grown from the vapor phase. Nonuniformities are observed particularly in the luminescence properties.

537.311.33:546.482.21 2328
Significance of the Electrical Contact in Investigations on CdS Single Crystals—W. M. Buttler and W. Muscheid. [*Ann. Phys* (*Leipzig*), vol. 14, pp. 215–219; February 15, 1954, and vol. 15, pp. 82–111; November 15, 1954.) The properties of ohmic and of nonohmic CdS photoresistors were investigated experimentally. Separate models are suggested for the metal-to-semiconductor contact in the two cases, one based on the assumption of a perfect contact, the other assuming a dependence of the contact on adsorption layers and surface states. Both models deviate from the Schottky-Mott model.

537.311.33:546.623.86 2329
Zone Melting of Aluminium Antimonide—H. A. Schell. (*Z. Metallkde*, vol. 46, pp. 58–61; January, 1955.) Experiments on the purification of AlSb were made using Pfann's technique. The method proved satisfactory for eliminating Cu, Fe, Mg, Si, Ca and Pb, and yielded material with improved electrical properties.

537.311.33:546.811–17 2330
Measurements of Electrical Conductivity and Magnetoresistance of Gray Tin Filaments—A. W. Ewald and E. E. Kohnke. (*Phys. Rev.*, vol. 97, pp. 607–613; February 1, 1955.) Measurements were made on filaments of pure material and of alloys containing Sb, As, In, Al or Zn prepared by the method described previously (*Phys. Rev.*, vol. 91, p. 244; July 1, 1953.) For 99.998 per cent pure Sn the activation energy is 0.082 ev and the conductivity at 0 degree C. is 2,090Ω^{-1}. cm^{-1}. The activation energy is increased by addition of impurities. Differences in the temperature variation of conductivity produced by the different types of impurity are indicated.

537.311.33:548.0 2331
Energy Levels of a Disordered Alloy—R. H. Parmenter. (*Phys. Rev.*, vol. 97, pp. 587–598; February 1, 1955.) A study is made of the one-electron energy levels of a disordered alloy, using perturbation theory. The method is relevant to problems of semiconductor crystals with imperfections. Some experimental verification is reported for a predicted tailing off of the density-of-states curve into a forbidden band.

537.311.33:621.396.822 2332
Current Noise in Semiconductors—W. Baumgartner and H. U. Thoma. (*Z. angew. Math. Phys.*, vol. 6, pp. 66–68; January 25, 1955.) The observed inverse-frequency spectral distribution of noise is explained in terms of electron trapping.

538.22:546.3-1-47-46-26 2333
The Structure and Properties of Some Ternary Alloys of Manganese, Zinc and Carbon—R. G. Butters and H. P. Myers. (*Phil. Mag.*, vol. 46, pp. 132–143; February, 1955.) The single-phase alloys prepared have an ordered structure of the perovskite type and show spontaneous magnetization at room temperature. The observed magnetic properties are exceptional and indicate a new type of ferrimagnetic behavior postulated by Néel (3159 of 1949) but not hitherto observed. Measured characteristics are presented.

538.221 2334
Possibility of the Observation of Exchange Resonance near a Ferrimagnetic Compensation Point—R. K. Wangsness. (*Phys. Rev.*, vol. 97, p. 831; February 1, 1955.) Theory for a system containing two sublattices indicates that resonance corresponding to the upper branch of the frequency curve should occur at a wavelength of the order of 0.1 mm. The possibility of observing such resonance is improved by working near the compensation point for angular momentum.

538.221 2335
Observation of Exchange Resonance near a Ferrimagnetic Compensation Point—T. R. McGuire. (*Phys. Rev.*, vol. 97, pp. 831–832; February 1, 1955.) Report of observations made under the conditions described by Wangsness (2334 above).

538.221 2336
Domains of Reverse Magnetization in Ferromagnetic Metals—T. G. Nilan and W. S. Paxton. (*Phys. Rev.*, vol. 97, pp. 834–835; February 1, 1955.) Calculations made by Goodenough (470 of March) are discussed in the light of studies of powder patterns in polycrystalline grain-oriented Si steels.

538.221 2337
Spontaneous Magnetogalvanism and Magnetization in Irreversible Ferronickels—A. L. Perrier and E. Ascher. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1066–1068; March 7, 1955.] Hall-effect and resistance measurements have been made on specimens containing about 70 per cent Fe and 30 per cent Ni, over a range of temperatures. Anomalies in the curves obtained are eliminated if, instead of the usual Hall emf, the current component normal to the longitudinal electric field and to the magnetic field or the magnetization is considered as the primary transverse effect.

538.221 2338
Fundamental Processes in the Magnetization of Alnico Permanent-Magnet Alloys—H. Fahlenbrach. (*Naturwiss.*, vol. 42, pp. 64–65; February, 1955.) Electron-microscope investigations of polycrystalline and single-crystal specimens of alnico-400 confirm the structural features observed by Nesbitt and Heidenreich (3456 of 1952) in heat-treated alloys.

538.221 2339
Dependence of Coercive Force on Thickness of Laminae of Fe-Si Alloy—Yu. P. Burdakova and V. V. Druzhinin. (*Zh. Tekh. Fiz.*, vol. 25, pp. 108–111; January, 1955.)

538.221:537.311.31:538.632 2340
Band Model for Hall Effect, Magnetization, and Resistivity of Magnetic Metals—E. M. Pugh. (*Phys. Rev.*, vol. 97, pp. 647–654; February 1, 1955.)

538.221:538.652 2341
The [110] Magnetostriction of Some Single Crystals of Iron and Silicon Iron—E. W. Lee. (*Proc. Phys. Soc.*, vol. 68, pp. 65–71; February 1, 1955.) The results of the previous investigation of magnetization curves (117 of 1954) are

used in magnetostriction calculations. The magnetostriction is found to depend on crystal width in a manner similar to that of the magnetization curves.

539.234:546.23 2342
X-Ray Investigation of Selenium Films obtained by Evaporation in Vacuum—D. N. Nasledov, V. A. Dorin and I. M. Dikina. (*Zh. Tekh. Fiz.*, vol. 25, pp. 29–38; January, 1955.) A detailed report is presented; one of the conclusions is that the electrical conductivity of the film depends on the temperature of the base during the evaporation process.

549.514.51:621.93 2343
A Novel Type of Saw for the Economical Cutting of Quartz Crystals or Other Materials—J. E. Thwaites and C. F. Sayers. (*P. O. Elect. Engr. Jour.*, vol. 47, part 4, pp. 233–235; January, 1955.)

621.3-761 2344
New Solution Ceramic Coatings—K. Rose. (*Materials and Methods*, vol. 41, pp. 107–108; February, 1955.) Heat- and corrosion-resistant films are formed on either metal or nonmetal surfaces by spraying on an aqueous solution of metallic salts. The surface is warmed sufficiently to drive off the water, leaving a tightly bonded oxide layer. Oxides of Zr and Cr are especially satisfactory; oxides of Ti, Ce and Mg and some phosphates, silicates, oxyhalides and metals have also been applied in this way. Such films may be useful for resistors and capacitors as well as for protective coatings.

621.315.615:537.311.3 2345
A Note concerning the Conductivity of Liquid Dielectrics—J. Hart and D. A. Simmons. (*Canad. Jour. Phys.*, vol. 33, pp. 54–57; February, 1955.) "Experiments are described which show that it is not permissible to ignore electrode effects in the calculation of the mobilities of ions in liquid dielectrics."

546.831 2346
Zirkonium—Seine Herstellung, Eigenschaften und Anwendungen in der Vacuumtechnik (Zirconium—its Production, Properties and Applications in Vacuum Technique) [Book Review]—W. Espe. Publishers: Winter'sche Verlagshandlung, Fuessen/Bayern, 1953, 174 pp. (*Electronic Eng.*, vol. 27, p. 232; May, 1955.) A comprehensive survey including tables of properties and a bibliography; of particular interest in relation to gettering technique.

MATHEMATICS

516 2347
Geometric Representation of the Gudermann Transformation—R. Cazenave. (*Ann. Télécommun.*, vol. 9, pp. 330–333; December, 1954.)

517.51 2348
The Theory of Generalized Functions—G. Temple. (*Proc. Roy. Soc. A*, vol. 228, pp. 175–190; February 22, 1955.) An introductory account of the construction and properties of generalized functions of real variables, so defined that any generalized function $f(x)$ possesses partial derivatives $D^pf(x)$ of all orders, and that if the sequence $f_n(x)$ converges to $f(x)$, then $D^pf_n(x)$ converges to $D^pf(x)$. Delta functions of Dirac are included in the theory. The representation of generalized functions by Fourier series and integrals is illustrated. Applications in vector analysis, potential theory and wave theory will be dealt with subsequently.

517.562 2349
Locus Diagrams of some Elementary Transcendental Functions—H. Wahl. (*Frequenz*, vol. 8, pp. 372–378; December, 1954.) Continuation of paper abstracted in 1423 of June.

517.65 2350
Method of Computation for Generalized Lommel Integrals—G. Coulmy. (*Ann. Télécommun.*, vol. 9, pp. 305–312; November, 1954.)

517.65:621.396.11 2351
A Table of a Function used in Radio-Propagation Theory—Horner. (*See* 2397.)

519.241.1:621.372.54 2352
A New Method of Determining Correlation Functions of Stationary Time Series—D. G. Lampard. (*Proc. IEE*, Part C, vol. 102, pp. 35–41; March, 1955. Digest, *ibid.*, Part III, vol. 101, pp. 343–346; September, 1954.) The method is based on an expansion of the correlation function in a suitable orthonormal system. The coefficients in the expansion are simply related to the convolution integrals which arise when the given time series are applied to linear filters whose impulse responses are members of this orthonormal system. A method of measuring these coefficients is proposed. Complex filters can be built up having an impulse response of the same form as the desired correlation function. Examples are presented of filters whose impulse responses are members of a Laguerre system. A practical autocorrelator based on such a filter is described and some results obtained with it are reported.

519.241.5 2353
On the Averaging of Data—S. S. Stevens. (*Science*, vol. 121, pp. 113–116; January 28, 1955.) A brief introductory discussion on when to use the mean, mode or median as measure of central tendency.

517 2354
Higher Transcendental Functions [Book Review]—Staff of the Bateman Project. Publishers: McGraw Hill, London, 1953, vol. 1, 302 pp., 52s., vol. 2, 396 pp., 60 s. [*Nature* (*London*), vol. 175, p. 317; February 19, 1955.] The first two volumes of a series. Thousands of formulas are presented, usually with some indication of origin and often with suggestions as to methods of proof.

MEASUREMENTS AND TEST GEAR

621.396.11.029.45:621.3.018.41(083.74) 2355
The Diurnal Carrier-Phase Variation of a 16-Kilocycle Transatlantic Signal—Pierce. (*See* 2404.)

621.317.3:621.314.632:546.289 2356
Recovery-Time Measurements on Point-Contact Germanium Diodes—T. E. Firle, M. E. McMahon and J. F. Roach. (PROC. IRE, vol. 43, pp. 603–607; May, 1955.) Discussion indicates the importance of using standard techniques and terminology.

621.317.3.012.3 2357
Phase-Linearity Nomograph—J. F. Sodaro. (*Electronics*, vol. 28, p. 178; April, 1955.) Nomogram relating delay time, phase shift and frequency for circuits of various types.

621.317.32:621.385.2 2358
Use of Diodes for Measurement of R.M.S. Voltage with Good Independence of Waveform—G. Bernardi. (*Ricerca Sci.*, vol. 25, pp. 34–48; January, 1955.) The important feature of the measurement circuit described is the inclusion of an improved voltage divider for supplying the diodes.

621.317.329:621.373.413 2359
Measurement of the Distribution of the Electromagnetic Field in a Resonant Cavity—A. Septier. (*Jour. Phys. Radium*, vol. 16, pp. 108–114; February, 1955.) See 1853 of 1954.

621.317.335.2 2360
The Precise Measurement of Capacitance—J. K. Webb and H. B. Wood. (*Proc. IEE*, Part C, vol. 102, pp. 3–12; March, 1955. Digest, *ibid.*, Part II, vol. 101, pp. 681–682; December, 1954.) The method described is designed to enable measurements to be made conveniently with a precision approaching that achieved by the National Physical Laboratory. A bridge network is used with inductively coupled ratio arms which can be switched into the form of either a comparison or a Wien bridge. By combined use of the comparison bridge with a 2:1 ratio and the Wien bridge with a 1:1 ratio the unknown capacitance in one arm can be determined in terms of the resistance in the other arm at a given frequency.

621.317.335.3.029.63/.64 2361
Phase and Amplitude Balance Methods for Permittivity Measurements between 4 and 50 cm—T. J. Buchanan and E. H. Grant. (*Brit. Jour. Appl. Phys.*, vol. 6, pp. 64–66; February, 1955.) The coaxial-line equipment described can be used to measure permittivities of liquids over a wide range of values; several alternative methods of procedure are given and two types of liquid cell are described. The apparatus is useful mainly for medium and high-loss liquids; for low-loss liquids cavity methods are preferred.

621.317.353.2 2362
Measurements of a Fundamental by [periodic] Contacting without using a Fundamental Filter—P. K. Hermann. (*Frequenz*, vol. 8, pp. 379–382; December, 1954.) Extension of the method in which harmonics are eliminated by appropriate choice of the contacting period.

621.317.374 2363
Loss-Angle Standards [fixed and variable]—H. Hoyer. (*Arch. Elektrotech.*, vol. 41, pp. 347–356; November 25, 1954.) Several types of fixed and variable capacitor-resistor loss-angle standards developed at the Physikalisch-Technische Bundesanstalt are described. The constructional details, method of use and experimental results are given. Standards for use at frequencies up to 10 kc are included.

621.317.374.027.3 2364
A Recording Loss-Angle Meter for High Voltages—K. B. Westendorf. (*Arch. Elektrotech.*, vol. 41, pp. 333–346; November 25, 1954.) A bridge-circuit instrument for use at voltages of the order of 10 kv at 50 cps is described.

621.317.38.029.63 2365
A Note concerning Forces and Torques on Spheroidal Bodies in Cavities—W. J. van de Lindt. (*Canad. Jour. Phys.*, vol. 33, pp. 113–117; February, 1955.) The method of determining microwave power from the force or torque on a ring in a cavity [1140 of 1954 (Kalra)] is modified by substitution of a metal sphere or spheroid for the ring. The forces acting on such a body are calculated from considerations of the energy in the cavity field.

621.317.411:621.318.134:621.372.413 2366
Circularly Polarized Cavities for Measurement of Tensor Permeabilities—E. G. Spencer, R. C. LeCraw and F. Reggia. (*Jour. Appl. Phys.*, vol. 26, pp. 354–355; March, 1955.) The determination of the tensor permeability of a ferrite by introducing a small sample into a cavity and observing the resonance characteristic is facilitated by exciting the cavity with a circularly rather than a linearly polarized wave.

621.317.42 2367
A Probe for the Study of Magnetic Fields—R. Birebent. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1064–1065; March 7, 1955.] The probe comprises a small coil fixed at one end of a tube and fed from an oscillator by leads passing through the tube, the other end of which carries a piezoelectric crystal. When the coil is exposed to a magnetic field, an alternating torsion is applied to the crystal, whose output is proportional to the field and to the coil current. Maximum sensitivity is obtained when the oscillator frequency is equal to the natural frequency of vibration of the probe, which was 1 kc in a particular model. The range of measurement is from a few oersteds to to some thousands of oersteds.

621.317.7:621.3.018.75 2368

A Transient Pulse Width and Pulse Amplitude Meter—F. Hart. (*Electronic Eng.*, vol. 27, pp. 192–197; May, 1955.) Two low-leakage capacitors are given charges proportional respectively to the amplitude and the duration of the pulse, and the voltages across the capacitors are measured with electrometer-type voltmeters before the charge leaks away. The particular instrument described was developed for observations on pulses originating from a mechanically vibrating tube.

621.317.7:621.385 2369

Circuit for the Determination of Contact Potentials and Electron Temperatures from Retarding Field Characteristics—S. Friedman and L. N. Heynick. (*Rev. Sci. Instr.*, vol. 26, pp. 17–19; January, 1955.) The design of the circuit is based on the retarding-field equation for parallel-plane diodes; errors involved in applying it to measurements on other tubes are discussed.

621.317.729.1 2370

Probe Impedance in the Electrolytic Tank—J. C. Burfoot. (*Brit. Jour. Appl. Phys.*, vol. 6, pp. 67–68; February, 1955.) The impedance is usually ascribed to capacitance or polarization effects, but in many cases the simple ionic conduction mechanism will give a useful estimate, so that the optimum size and shape may be calculated. Theoretical results are supported by measurements.

621.317.733:621.375.2 2371

A Tuned Detector-Amplifier for Power-Frequency Measurements—W. K. Clothier and W. E. Smith. (*Jour. Sci. Instr.*, vol. 32, pp. 67–70; February, 1955.) A detector-amplifier for bridge measurements is tuned by means of a two-section LC filter, the inductors of which are varied simultaneously by dc in control windings on their mumetal cores. When used with suitable tuned input transformers it can detect power as low as 10^{-19} w.

621.317.75:[621-526+621.372.5 2372

A Harmonic Response Plotter—Z. Czajkowski. (*Electronic Eng.*, vol. 27, pp. 207–211; May, 1955.) The instrument described gives simultaneously the frequency response of a system and its first derived plot. It was designed for testing servo systems, but is useful for determining the characteristics of any network over the frequency range 0.2–104 cps. An electromechanical analog computer system is used incorporating six servomechanisms.

621.317.75:[621-526+621.372.5 2373

Servo Analyzer for Wide-Range Testing—F. E. Dickey. (*Electronics*, vol. 28, pp. 172–175; April, 1955.) A phase-shift oscillator covering the range 0.25–300 cps provides a drive voltage for the servo unit under test and a variable two-phase output for analysis purposes. It may also be used to measure the characteristics of filters etc. The output can be modulated to supply a 60- or 400-cps suppressed-carrier AM signal.

621.317.75:621-526 2374

Transient Measurement of Feedback Control Systems—F. H. Ferguson and C. H. Looney. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 110–114; May, 1953.) Description of a transient-response indicator and recording system.

621.317.75:621.373.4.029.62 2375

Wobbulator Adaptor for Band III—G. H. Leonard. (*Wireless World*, vol. 61, pp. 283–287; June, 1955.) Full details are given of a circuit designed to operate with an existing band-I frequency-sweep oscillator for alignment of tuning units. Third harmonics of the 57–73-mc range are derived in a tripler stage and amplified, the response of the amplifier peaking somewhat below the center of the pass band to discriminate against fourth harmonics. An over-all flat response is achieved by an agc system based on detecting any modulation at the sweep frequency, 25 cps, in the rf output. Marker, attenuator and cro display systems are described.

621.317.755 2376

The Development and Design of Direct-Coupled Cathode-Ray Oscilloscopes for Industry and Research—M. J. Goddard. (*Jour. Brit. IRE*, vol. 15, pp. 179–197; April, 1955.)

621.317.755 2377

Study and Realization of a Four-Beam Cathode-Ray Oscillograph for High-Voltage Operation—J. Ollé. (*Rev. Gén. élect.*, vol. 64, pp. 104–108; February, 1955.) A fuller account of the oscillograph previously described by Fert et al. (1624 of 1954).

621.317.755:621.318.57 2378

Trigger Adapter for Transient Oscillograms—L. Fleming. (*Electronics*, vol. 28 pp. 159–161; April, 1955.) An auxiliary unit operated by the transient to be displayed provides single sweep and unblanking of the cro beam. Sweep times of 2–5 ms are used.

621.317.755:621.385.029.6 2379

Velocity Spectrography of Electron Dynamics in a Traveling Field—O. T. Purl and H. M. VonFoerster. (*Jour. Appl. Phys.*, vol. 26, pp. 351–353; March, 1955.) Velocity variations of electrons at a selected point along a beam are studied by means of a modification of an arrangement described by Bloom and VonFoerster (218 of February), one of the Lecher-wire deflection systems being replaced by a pair of curved deflection plates and the fluorescent screen being correspondingly offset. Use of this device for investigating the motion of electrons at the input of a traveling-wave tube is discussed.

621.317.772 2380

A Direct-Indicating Phase Meter—A. van Weel. (*Jour. Brit. IRE*, vol. 15, pp. 143–152; March, 1955.) Description of a complete instrument based on principles discussed previously (3659 of 1953).

621.317.78.029.6 2381

A Microwave Microcalorimeter—A. C. Macpherson and D. M. Kerns. (*Rev. Sci. Instr.*, vol. 26, pp. 27–33; January, 1955.) A bolometer instrument developed at the National Bureau of Standards measures microwave power accurately to within 1 per cent at milliwatt levels. The bolometer mount assembly serves as a matched waveguide termination.

621.373.421.13:621.3.018.41(083.74) 2382

A Simple Circuit for Frequency Standards employing Overtone Crystals—E. P. Felch and J. O. Israel. (PROC. IRE, vol. 43, pp. 596–603; May, 1955.) An oscillator circuit with frequency stable to within one part in 10^9 over periods of several hours is obtained using a single-stage class-A amplifier with Π-section feedback network and a quartz crystal operating at its fifth overtone, at 5 mc [see 3481 of 1952 (Warner)]. Undesired frequencies are suppressed by connecting a resistor in parallel with the crystal.

621.385.001.4:531.768 2383

Vibration Generator for Electron Tubes—(*Elec. Jour.* vol. 154, p. 350; February 4, 1955.) Electrodynamic apparatus developed at the National Bureau of Standards for testing subminiature tubes uses a $BaTiO_3$ accelerometer capable of measuring vibration frequencies up to 20 kc at acceleration levels of 0.2–10,000 g.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.222–8 2384

Photoelectric Measurement of Weak Acoustic Birefringence—J. Badoz. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 1062–1064; March 7, 1955.] An ultrasonic beam modulated at low frequency is passed through a liquid, thus periodically producing birefringence. Polarized light passes through the liquid and is received by a photomultiplier, the intensity of the received light varying with the ultrasonic variations. The photomultiplier output is connected to a selective amplifier followed by a detector. Differences of optical path down to $10^{-4}\lambda$ can thus be measured.

621.316.7 2385

A Design Philosophy for Man-Machine Control Systems—M. Hoberman. (PROC. IRE, vol. 43, p. 623; May, 1955.) Comment on 822 of March (Birmingham and Taylor), suggesting that animals could replace humans as machine minders in some cases.

621.37/.38]786.2 2386

New Electronic Piano—B. F. Miessner. (*Radio-Electronics*, vol. 26, pp. 64–66, 68; February, 1955.) Description of a keyboard instrument in which the strings are replaced by hammer-struck fixed-free reeds capacitively associated with fixed electrodes to control the frequency of an oscillator, producing a FM signal which is detected by a discriminator.

621.373.43:616–7 2387

Application of Pulse Circuits in Medicine—I. Abrikosov. [*Radio (Moscow)*, no. 12, pp. 43–45; December, 1954.] Basic circuits are shown of apparatus for diagnosis and therapy. The pulse shapes and repetition rates most commonly used are indicated.

621.383.2:621.317.087:534.213–8 2388

A Method of Ultrasonoscopy using an Electron-Optical Image Converter—M. von Ardenne. (*Nachr Tech.*, vol. 5, pp. 49–51; February, 1955.) The ultrasonic-wave intensity distribution produced by a selectively absorbing test object is converted into a corresponding temperature distribution at an absorbing layer on the image converter photocathode. For maximum sensitivity the cathode work function should be low; a suggested material is [Ag]-Cs_2O on a CsAg base, which has a work function of about 1 ev. The sketch of the suggested arrangement includes notes on materials and dimensions; the calculated threshold ultrasonic energy density, when using a photographic method of recording the image, is about 0.04 w/cm². References to literature on the subject are tabulated, classified according to the method used, and the operation of the converters is outlined.

621.383.2:778.37 2389

Two New Image-Converter Tubes for Scientific Uses, "Valvo 18120" and "Valvo 18130"—F. Cubasch. (*Elektronische Rundschau*, vol. 9, pp. 5–11; January, 1955.) A diode- and a triode-type image converter are described. Applications noted include use in a high-speed camera; a circuit diagram is given and results obtained are illustrated. Exposure times shorter than 10^{-7} seconds are attainable. Methods of increasing the repetition rate to about 200,000 exposures/second are mentioned.

621.384.622 2390

Linear-Accelerator Issue—(*Rev. Sci. Instr.*, vol. 26, pp. 111–231; February, 1955.) Detailed accounts are included of the Berkeley accelerator, which produces 31.5-Mev protons, and of the Stanford Mark-II and Mark-III traveling-wave electron accelerators.

621.385.833 2391

Electron Optics of Cylindrical Systems having a Plane of Symmetry: Part 1—First-Order Approximation—M. Laudet. (*Jour. Phys. Radium*, vol. 16, pp. 118–124; February,

1955.) Expressions for the electron trajectories are derived, using Lagrange's method.

621.385.833:537.533 2392

Use of Space Charge in Electron Optics—E. A. Ash. (*Jour. Appl. Phys.*, vol. 26, pp. 327–330; March, 1955.) "The electron optical properties of a cylindrical space charge cloud are derived. The possibility of achieving a system free from either spherical or chromatic aberration by combining a space charge lens with a space charge free converging lens is examined. It is concluded that the achromatization of a thin electric or magnetic lens is possible only if the resultant action of the combination is divergent. The correction of the spherical aberration of a high quality lens, such as an electron microscope objective, is shown to be impossible on account of electron interaction effects."

621.387.424 2393

Geometric Efficiency of Cylindrical [G-M] Counters—J. Goldemberg. (*Rev. Sci. Instr.*, vol. 26, pp. 41–44; January, 1955.)

621.387.424 2394

Spontaneous Discharges of Thermionic Origin in Geiger Counters—R. Meunier, M. Bonpas and J. P. Legrand. (*Jour. Phys. Radium*, vol. 16, pp. 145–148; February, 1955.)

621.387.424 2395

Geiger Counters with Tr_2O Vapour Filling—P. Meunier, M. Bonpas and J. P. Legrand. (*Jour. Phys. Radium*, vol. 16. pp. 148–151; February, 1955.)

621.397.3 2396

Image Processing—Kovásznay and Joseph. (*See* 2445.)

PROPAGATION OF WAVES

621.396.11:517.65 2397

A Table of a Function used in Radio-Propagation Theory—F. Horner. (*Proc. IEE*, Part C, vol. 102, pp. 134–137; March, 1955. Digest, *ibid.*, Part B, vol. 102, p. 400; May, 1955.) "Real and imaginary components of the complex integral $1+2j\sqrt{(\omega)}\epsilon^{-\omega}\int_{-j\sqrt{\omega}}^{\infty}\epsilon^{-x^2}dx$ are tabulated as a function of ω. The Tables are prepared for real components of ω in the range 0 to 10, associated with imaginary components in the range -10 to 0."

621.396.11:537.56 2398

Theory of Coupling of the Ordinary and Extraordinary Electromagnetic Waves in an Inhomogeneous Anisotropic Plasma and Conditions for Reflection. Applications to the Ionosphere—R. Jancel and T. Kahan. (*Jour. Phys. Radium*, vol. 16, pp. 136–145; February, 1955.) Results of previous work [1030, 2196 and 2364 of 1954] are applied. Using the BKW approximation, expressions for the refractive index of the medium and the conditions for reflection both for longitudinal and transverse propagation are developed. Coupling is then considered, the four-sheeted Riemann surface representing the refractive-index equation giving the various forms of coupling between the ordinary and extraordinary rays. This surface is studied in detail for the particular conditions of the ionosphere.

621.396.11:551.510.52 2399

Cheyenne Mountain Tropospheric Propagation Experiments—A. P. Barsis, J. W. Herbstreit and K. O. Hornberg. (*NBS Circulars*, no. 554, 39 pp; January 3, 1955.) Experiments were started in 1950 on propagation over simulated air-to-ground communication paths, using cw transmitters on five frequencies in the range 92–1 046 mc, receivers continuously recording field strength at four locations within 226 miles of the mountain with additional provision for reception at distances of 393 and 617 miles, and an extensive meteorological installation at Haswell, Colo. A detailed description is given of the transmitting and receiving equipment. Sample results lend support to Booker and Gordon's theory of scattering elaborated by Staras (804 of 1953). See also 2499 of 1954 (Bean) and *1955 IRE Convention Record*, Part 2, pp. 85–93 (Herbstreit et al.).

621.396.11:551.510.535 2400

An Ionospheric Attenuation Equivalence Theorem—E. V. Appleton and W. J. G. Beynon. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 141–148; March, 1955.) Partial reflection and scattering in the ionosphere may be important sources of attenuation of radio waves. For frequencies well above the equivalent critical frequency the following theorem relating oblique-incidence to f vertical-incidence phenomena is stated: $[\rho]_{i_0}^{f}=\lambda[\rho]_0^{f\cos i_0}$ where ρ is the fractional attenuation, f the frequency and i_0 the angle of incidence; $\lambda=1$ or $\cos 2\, i_0$, according as the electric vector is perpendicular to the plane of propagation or in the plane.

621.396.11:551.510.535 2401

Radio Disturbance Warnings in Japan—T. Obayashi. [*Jour. Radio. Res. Labs* (*Japan*), vol. 1, pp. 55–61; September, 1954.] The system operated by the Hiraiso observatory is described. Short-term forecasts are broadcast on 4 mc and 8 mc from the standard-frequency station JJY. Weekly and monthly forecasts and special services for operators are provided.

621.396.11.029.45:551.510.535 2402

The Numerical Solution of Differential Equations governing Reflexion of Long Radio Waves from the Ionosphere—K. G. Budden. (*Proc. Roy. Soc. A*, vol. 227, pp. 516–537; February 8, 1955.) Two methods which have been used with the EDSAC are described. "In the first method the first-order simultaneous equations, derived from Maxwell's equations and the constitutive relations for the ionosphere, are integrated by a step-by-step process proceeding downwards. The integrations are started from properly chosen initial solutions. From the resulting field variables at the bottom of the ionosphere a reflexion coefficient matrix **R** is derived, whose elements include the familiar reflexion coefficients. Two integrations are needed for each derivation of the elements of **R**. For the second method, it is shown that the formulae for **R** for a level below the ionosphere can be applied also within the ionized medium, and define a more general matrix variable whose elements are the dependent variables in a new set of differential equations. These are integrated by a step-by-step process as in the first method. The solution obtained below the ionosphere gives the required set of reflexion coefficients without further calculation. Only one integration is required for each derivation. The equations are given in full for certain important special cases."

621.396.11.029.45:551.510.535 2403

The Ionospheric Propagation of Radio Waves of Frequency 16 kc/s over Short Distances—T. W. Straker. (*Proc. IEE*, Part C, vol. 102, pp. 122–123; March, 1955. Digest, *ibid.*, Part B, vol. 102, pp. 396–399; May, 1955.) Report of experiments made during the period March, 1948–October, 1949, in continuation of the investigations described by Budden *et al.* (*Proc. Roy. Soc. A*, vol. 171, pp. 188–214; May 19, 1939) on signals transmitted from Rugby to Cambridge; amplitude and phase were measured separately. Conclusions reached previously concerning diurnal and seasonal variations of the reflection height are confirmed. Marked seasonal variation of the amplitude of the downcoming wave was observed, the variations being different for daytime and night-time. During summer the amplitude was closely controlled by the sun, but was not related in any simple way to the zenith angle. The sunrise effect for amplitude occurred about an hour earlier than for phase. Subsidiary variations observed are consistent with reflection taking place at two levels. Abnormal variations of reflection height occurred during and after magnetic storms.

621.396.11.029.45:621.3.018.41(083.74) 2404

The Diurnal Carrier-Phase Variation of a 16-Kilocycle Transatlantic Signal—J. A. Pierce. (Proc. IRE, vol. 43, pp. 584–588; May, 1955.) Measurements at Cambridge, Mass., on signals from GBR are reported. The diurnal variation in apparent transmission time is about 40 μs; this is presumed to be due to a variation of 10–12 km in h'. The degree of phase stability of the received signal is such as to permit intercontinental comparison of frequencies with a precision of at least 1 part in 10^{10}. On this basis, a radio navigation system could theoretically be accurate to within about a mile at a range of several thousand miles. The importance of such stability for narrow-band reception is discussed. See also 1425 of June (Pierce et al.).

621.396.11.029.6 2405

Tropospheric Scatter Propagation—(*Wireless World*, vol. 61, pp. 253–254; June, 1955.) Experiments carried out in America at frequencies in the lower uhf band, using 10-kw transmitters with antenna reflectors 60 feet in diameter, show that wide-band transmissions over a range of 200 miles are feasible. The system will probably be used to supplement the present radio-link system. Other experiments in progress at Syracuse are designed to determine the best modulation system and the diurnal and seasonal variation in reception. See also 2811 of 1951 (Straiton et al.).

621.396.11.029.6 2406

Large Reduction of V.H.F. Transmission Loss and Fading by the Presence of a Mountain Obstacle in Beyond-Line-of-Sight Paths—J. H. Crysdale, F. H. Dickson, J. J. Egli, J. W. Herbstreit and G. S. Wickizer. (Proc. IRE, vol. 43, pp. 627–628; May, 1955.) Comment on 3396 of 1953 and authors' reply.

621.396.11.029.62 2407

V.H.F. Field Strength far beyond the Radio Horizon—T. F. Rogers. (Proc. IRE, vol. 43, p. 623; May, 1955.) Measurements on a 220-mc pulsed signal are reported; an airborne receiver was used, at an altitude of 500 feet, at distances up to 420 miles. Differences from Megaw's results (973 of 1951) are noted. The results appear to be reasonably consistent with a d^{-7} law of variation of field strength.

621.396.11.029.62 2408

Meteorological Consideration on the V.H.F. Propagation between Inubo and Hiraiso—K. Hirao and H. Maruyama. [*Jour. Radio. Res. Labs* (*Japan*), vol. 1, pp. 31–40; September, 1954.] Results of propagation tests during the period June–August, 1952 at 153 mc over a 78-km sea path are analyzed with reference to aerological data. A sudden 20-db increase of field strength lasting up to several hours occurred frequently during both baiu and summer seasons. These increases are related to meteorological conditions giving rise to dry, warm air over the sea surface.

621.396.11.029.62:551.510.535 2409

Sporadic-E Propagation—N. C. Gerson (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 113–116; March, 1955.) Reports of two-way communication by amateurs during the period 1949–1952, mainly in summer, on a frequency of 50 mc, are analyzed. The results indicate that the modal station separation was 1 490 km, 2,800 km and about 4,200 km for single-, double- and triple-hop communication respectively. E_s clouds having an effective diameter of about 925 km appeared to be fairly prevalent.

621.396.11.029.64:535.4 2410

The Reflection of Electromagnetic Waves from a Rough Surface—H. Davies. (*Proc. IEE*, Part C, vol. 102, p. 148; March, 1955.) Discussion on 537 of March.

621.396.81.029.53 2411

Measurement of Loran Waves—Y. Aono, T. Kobayashi, C. Ouchi and C. Nemoto. [*Jour. Radio Res. Labs* (*Japan*), vol. 1, pp. 1–16; September, 1954.] In order to develop a reliable system for predicting field strength in the mf band, measurements were made at Hiraiso of the field strength of 1.85-mc and 1.95-mc loran transmissions from 12 stations in the western Pacific Ocean between 200 and 4,000 km away. Results are analyzed and characteristic curves are derived for diurnal variation factor κ and absorption factor Γ_m. These are used in constructing nomograms (a) relating κ to local time, month, and latitude of transmitting and receiving stations, and (b) relating Γ_2 (deviative absorbtion factor), critical frequency, muf, operating frequency and distance.

RECEPTION

621.39:534.78 2412

The Effect of Severe Amplitude Limitation on Certain Types of Random Signal: a Clue to the Intelligibility of "Infinitely" Clipped Speech—J. M. C. Dukes. (*Proc. IEE*, Part C, vol. 102, pp. 88–97; March, 1955. Digest, *ibid.*, Part B, vol. 102, pp. 261–263; March, 1955.) The energy spectrum of a random waveform is compared with that of the waveform after limiting and differentiation. With certain signals of common occurrence the average energy distribution is not much affected by the limiting, and the phase relation between the waveforms is highly coherent. This result explains the observed high intelligibility of clipped speech [1537 of 1948 (Licklider and Pollack)].

621.396.621+621.37]049.75 2413

Investigations of Laboratory Production of Printed Circuits for Communication Equipment—Götze. (*See* 2223.)

621.396.621.57:621.314.7 2414

Crystal Valve Receivers—A. Stead. (*RSGB Bull.*, vol. 30, pp. 378–380; February, 1955.) The design of simple "straight" receivers employing point-contact and junction-type transistors is described and complete circuit diagrams are given. The receivers worked satisfactorily at frequencies up to 2 mc.

621.396.822:621.396.62 2415

The Response of a Nonlinear System to Random Noise—W. E. Thomson. (*Proc. IEE*, Part C, vol. 102, pp. 46–48; March, 1955. Digest, *ibid.*, Part III, vol. 101, p. 407; November, 1954.) A simplified form is derived of a formula obtained previously [see e.g. 3238 of 1948 (Middleton)] for the autocorrelation function of the output of a nonlinear system for the case where the output/input relation is expressed as a power series. The new formula emphasizes that intermodulation products of different orders are uncorrelated while contributions to the intermodulation product of a given order from different terms of the power series are completely correlated. If the output is expressed as a series of Hermite polynomial functions of the input, each term of the series gives rise to intermodulation products of one order only.

621.396.823 2416

High-Frequency Interference from Electric-Fence Equipment—M. Haag. (*Elektrotech. Z.*, *Edn. A*, vol. 76, pp. 120–124; February 1, 1955.) Electric fences considered are operated with voltage pulses of up to 5 kv, maximum pulse charge of 2.5 ma.s, pulse duration 0.1 second and pulse intervals of at least 0.75 second. Radio interference occurs mainly in the long-, medium- and short-wave bands; usw interference has not so far been reported. Suitable measurement circuits and suppressors are discussed and both impedance/frequency and interference/frequency curves are shown.

621.396.823:537.523.3 2417

Radio Interference from Direct-Voltage Corona in a Coaxial Cylindrical Field—H. Heindl. (*Arch. elekt. Übertragung*, vol. 9, pp. 93–98; February, 1955.) Report of a laboratory investigation intended as preliminary to a study of the effects produced by ac lines. The rf interference calculated from a consideration of the corona discharge mechanism is in good agreement with the observations. Comparison is made also with results obtained by other workers in various countries.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11-20 2418

Extent of Redundancy in the Speech and its Importance in Long-Distance Telephony—W. Endres. (*Fernmeldetech. Z.*, vol. 8, pp. 89–93; February, 1955.) According to communication theory, the time interval during which the frequency/time characteristic of a sound remains constant is considered to be redundant. Visible speech diagrams indicate that the redundancy in spoken English is about 50 per cent; a 2:1 time-compression should therefore be feasible.

621.395.44:621.315.052.63 2419

Telecommunication Equipment for Power Systems: Developments and Application in Sweden—U. Hecht, S. Rodhe and H. J. B. Nevitt. (*Trans. Amer. IEE*, Part III, *Power Apparatus and Systems*, vol. 72, pp. 961–968; October, 1953. Discussion, pp. 968–969.) See also 1148 of May (Rathsman et al.)

621.395.44:621.315.052.63:621.395.822 2420

A Study of Carrier-Frequency Noise on Power Lines: Part 3—Interpretation of Field Measurements—J. D. Moynihan and B. J. Sparlin. (*Trans. Amer. IEE*, Part III, *Power Apparatus and Systems*, vol. 72, pp. 573–580; June, 1953. Discussion, pp. 580–581. Corrections for bandwidth and impedance differences to be considered in correlating test data are discussed. A standard reference bandwidth of 1 kc is suggested. Parts 1 and 2: 3705 of 1953 (Cheek and Moynihan).

621.395.741:621.3.018.78 2421

Law of Addition of Distortion Voltages in Long-Distance Communication Systems—K. Steinbuch and H. Marko. (*Fernmeldetech. Z.*, vol. 8, pp. 71–78; February, 1955.) The law of addition of interference voltages due to thermal noise and to the nonlinear distortion in the amplifiers is derived for cascaded amplifiers with given phase characteristics. With sufficiently curved characteristics arithmetical addition is partly prevented, but the maximum obtainable gain is of the order of only 1.3 *n*. Such a phase characteristic can be obtained by means of a suitable all-pass filter in the system.

621.396.4:621.376.222 2422

Study of the Balanced Valve Modulator for S.S.B. Radio Communication—G. Bronzi. (*Alta Frequenza*, vol. 23, pp. 335–356; December, 1954.) Operation of conventional balanced bridge circuits is analyzed. If the amplitudes of the modulation and carrier signals, the static tube characteristics and the load are known, the maximum bridge output voltage can be determined graphically. Experiments confirm the validity of the method.

621.396.65:621.396.41 2423

Trieste - Venezia - Bologna - Verona - Milano Radio-Link Network—L. Bernardi. (*Poste e Telecomun.* vol. 23, pp. 61–67; February, 1955.) Extensions to the Mestre-Trieste multichannel microwave communication link (3109 of 1953) are described.

621.396.712.3:534.86 2424

On the Structural and Room Acoustics of the Multipurpose Studio Unit at Broadcasting House, Hamburg—Venzke. (*See* 2180.)

621.396.97+621.397.74 2425

Present State of U.S.W. Sound and Television Broadcasting Coverage [in Germany]—R. Gressmann. (*Funk-Technik*, vol. 10, pp. 31–34; January, 1955.)

SUBSIDIARY APPARATUS

621–526 2426

Asymmetrical Servomechanisms—J. Loeb and J. D. Lebel. (*Ann. Télécommun.*, vol. 9, pp. 282–286; October, 1954.) Servomechanisms in which the nonlinear element has an input/output characteristic asymmetric with respect to the origin are considered. Even if the input signal is a pure sine wave, the output will have a dc component that cannot be filtered out by the linear element. The general feedback-loop equations are reformulated and the new conditions of stability considered. A temperature-control device is treated as an example.

621–526 2427

Feedback Control Systems—(*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72; 1953.) The following papers are included:

"Quick Methods for evaluating the Closed-Loop Poles of Feedback Control Systems," —G. Biernson (pp. 53–68. Discussion, pp. 68–70).

"Correlation Between Frequency and Transient Responses of Feedback Control Systems,"—Y. Chu (pp. 81–92).

"Coulomb Friction in Feedback Control Systems,"—V. B. Haas, Jr. (pp. 119–123. Discussion, pp. 123–126).

"Describing Function Method of Servomechanism Analysis applied to Most Commonly Encountered Nonlinearities,"—H. D. Greif (pp. 243–248).

"Optimization of Nonlinear Control Systems by means of Nonlinear Feedbacks,"—R. S. Neiswander and R. H. MacNeal (pp. 262–270. Discussion, pp. 270–272).

"A Relative Damping Criterion for Linear Systems,"—J. F. Koenig (pp. 291–294. Discussion, pp. 294–295).

"Errors in Relay Servo Systems,"—L. F. Kazda (pp. 323–328. Discussion, p. 328).

"Effects of Friction in an Optimum Relay Servomechanism,"—T. M. Stout (pp. 329–335. Discussion, pp. 335–336).

"Backlash in a Velocity Lag Servomechanism," —N. B. Nichols (pp. 462–467).

"Determination of the Maximum Modulus, or of the Specified Gain, of a Servomechanism by Complex-Variable Differentiation,"—T. J. Higgins and C. M. Siegel (pp. 467–468. Discussion, pp. 468–469).

Abstracts of other papers appear separately.

621–526 2428

Some Saturation Phenomena in Servomechanisms with Emphasis on the Tachometer Stabilized Systems—E. Levinson. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 1–9; March, 1953.) Analytical methods of dealing with frequency response and transient response are described and various saturation effects are explained.

621–526 2429

Relative Stability of Closed-Loop Systems—M. J. Kirby and D. C. Beaumariage. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 22–43; March, 1953.) The equivalence of a transfer-function plot and a plot of the Laplace transform of transient response is illustrated and the application of transform plots in the analysis of linear systems is described.

621–526 2430

A Differential-Analyzer Study of Certain

Nonlinearly Damped Servomechanisms—R. R. Caldwell and V. C. Rideout. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 165–169; July, 1953. Discussion, pp. 169–170.) Advantages of the system described by Lewis (1186 of 1954) over corresponding linear systems are illustrated and modifications to improve its operation are suggested.

621–526 2431

Limiting in Feedback Control Systems— R. J. Kochenburger. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 180–192; July, 1953. Discussion, pp. 192–194.) Study of the effect of limiting on system performance in terms of the frequency response.

621–526 2432

Open-Loop Frequency Response Method for Nonlinear Servomechanisms—R. L. Cosgriff. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 222–225; September, 1953.) A graphical representation similar to a Nyquist diagram is developed for indicating frequency response characteristics and for the synthesis of linear filters.

621–526 2433

The Synthesis of "Optimum" Transient Response: Criteria and Standard Forms—D. Graham and R. C. Lathrop. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 273–286; November, 1953. Discussion, pp. 286–288.) "Eight mathematical criteria for optimum transient responses are critically examined, and the clear superiority of the minimum integral of time-multiplied absolute-value of error is demonstrated. The application of this criterion results in the selection of standard forms, which are presented in tables." See also 268 of February.

621–526 2434

Approximation of Transient Response from Frequency Response Data—C. H. Dawson. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 289–291; November, 1953.) A relatively short method is described involving an 18-point graphical integration which gives accurate results for linear feedback systems with a closed-loop transfer function of fifth order or below.

621–526:016 2435

Bibliography on Feedback Control—(*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 430–462; January, 1954.) A classified list of references up to 1952 relating to the theory and application of feedback systems. A list of relevant periodicals is also given.

621–526:621.3.066.6 2436

Control of Metal Build-up in Minimum Pressure Sensitive Contact Systems [for servomechanisms]—J. P. Dallas and T. R. Stuelpnagel. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 398–403; January, 1954.) A review of designs aimed at reducing spire-type metal transfer on contacts with gaps of 0.001–0.005 inch and operating on pressures down to 0.01 g.

621.311.6:621.373.4 2437

Highly Regulated R.F. Voltage Supply—L. G. Sloan, R. W. Raible and M. K. Testerman. (*Electronics*, vol. 28, pp. 192–200; April, 1955.) A rf output voltage of 5 v rms maximum, with low harmonic content, is obtained by using a crystal-controlled pentode oscillator at the fundamental frequency, with positive feedback. Regulation of the output is achieved by means of a negative-feedback loop incorporating a high-gain dc amplifier which supplies the oscillator screen-grid voltage.

621.314.63:546.28 2438

Silicon Power Rectifiers for AC Line Operation—G. Rudenberg. (*Electronics*, vol. 28, pp. 146–149; April, 1955.) Used as half-wave rectifiers, with capacitor-input filters, typical alloy-junction units rated for 135-v maximum rms input give 150 ma output at 125 degrees C. A typical full-wave rectifier with choke-input filter will deliver 800 ma at 125 degrees C. or 1.5 a at room temperature. Heat-dissipating mountings are discussed.

621.314.67:621.372.54 2439

Rectifier-Filter Characteristics—F. G. Heymann. (*Wireless Eng.*, vol. 32, pp. 147–154; June, 1955.) Approximate calculations are described for the design of half-wave rectifiers with capacitor-input filters and of full-wave rectifiers with capacitor-choke-input filters. The method gives results within about 5 per cent of measured values.

621.316.7.012 2440

Determination of Frequency Characteristics of Automatic Control Systems using Mikhailov's Curves—O. P. Demchenko. [*Compt. Rend. Acad. Sci. URSS*, vol. 100, pp. 693–696; February 1, 1955. In Russian.]

621.316.93:621.396.933 2441

Aircraft Protection from Thunderstorm Discharges to Antennas—J. M. Bryant, M.M. Newman and J. D. Robb. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 248–252; September, 1953. Discussion, pp. 253–254.) See 1586 of 1954.

621.316.993.08 2442

The Effect of Reactive Components in the Measurement of Grounding Circuits—L. H. Harrison. (*Trans. Amer. IEE*, Part II, *Applications and Industry*, vol. 72, pp. 340–343; November, 1953. Discussion, pp. 343–345.) Discussion of results of measurements of earth resistance using ac in the range 30 cps–200 kc, and dc.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2:621.397.8 2443

Some Typical Characteristics of Frequency-Modulated Television Transmissions—G. Brühl. (*Arch. elekt. Übertragung*, vol. 9, pp. 63–68; February, 1955.) Control of picture quality in the course of international television exchanges is considered. The distortion of the video signal by multiple reflections in bf and IF cables is discussed; the frequency swing is the determining factor. Methods of specifying the permissible deterioration and of measuring the relevant system transmission properties are indicated.

621.397.24 2444

The London-South Wales Television Link—J. C. D. Bell (*GEC Telecommun.*, no. 18, pp. 4–13; December, 1953.) Simultaneous two-way transmission is provided on the 165-mile London-Wenvoe coaxial link. The over-all gain/frequency characteristic is flat within ±0.65 db and the delay/frequency characteristic is flat within ±0.075 μs over the range 500 kc–3.8 mc. The rms noise level is 58 db below carrier level for the London-Bristol section.

621.397.3 2445

Image Processing—L. S. G. Kovásznay and H. M. Joseph. (Proc. IRE, vol. 43, pp. 560–570; May, 1955.) Mathematical theory is presented and extensions and applications are discussed of the techniques described previously (1541 of 1954).

621.397.335.001.4 2446

Aligning TV Receivers by Pulse-Cross Display—H. E. Thomas. (*Electronics*, vol. 28, pp. 184–192; April, 1955.) Synchronization faults in a television system may be analyzed by presentation on the picture tube screen of the synchronizing and blanking portions of the signal, by simple adjustment of the receiver controls. See also 1125 of 1952 (Launer).

621.397.5 2447

Television and Modern Information Theory—F. Schröter. (*Arch. elekt. Übertragung*, vol. 9, pp. 1–7; January, 1955.) See 556 of March.

621.397.5(083.7) 2448

I.R.E. Standards on Television: Definitions of Television Signal Measurement Terms, 1955—(Proc. IRE, vol. 43, pp. 619–622; Standard 55 IRE 23 S1.

621.397.6 2449

Possible Methods of Equalization in Vestigial Sideband Transmission in Television—H. J. Griese. (*Fernmeldetech. Z.*, vol. 8, pp. 94–103; February, 1955.) A survey with particular reference to the method of improving transient response described by Ruston (3253 of 1952). Results of an experimental investigation of transient response on two types of receiver are presented graphically and a standard television-receiver response is proposed, combined with the application of group delay pre-emphasis at the transmitter.

621.397.6:778.5:621.395.625.3 2450

Methods of Picture-Synchronized Sound Recording in Television—K. E. Gondesen. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 6, nos. 11/12, pp. 237–242; 1954.) 16-mm film with magnetic sound track is preferred. The principles of the single-stripe method and three variants of the double-stripe method, (a) with perforated film, (b) with a pilot frequency and (c) without special synchronization, are outlined. The equipment and operation of these methods are described and applications of these and photoelectric recording techniques are tabulated. See also 1802 of July (Lauer and Schulze).

621.397.611.2 2451

Properties and Applications of Television Camera Tubes with Photoconductive Targets—W. Heimann. (*Arch. elekt. Übertragung*, vol. 9, pp. 13–19; January, 1955.) The construction and operation of vidicon-type tubes are discussed with particular reference to the development of a German tube, the resistron. A method is described for measuring the time delay of the target response in normal operation, in which a vertical bar pattern is scanned horizontally across the target. The picture quality attained is illustrated by photographs.

621.397.62+621.396.67 2452

New Television Receivers and Aerials [in Western Germany]—W. W. Diefenbach. [*Funk-Technik (Berlin)*, vol. 10, pp. 88–91; February, 1955.] A brief survey of commercially available equipment.

621.397.62:535.88 2453

M.E.P. 55 Projector—A. V. J. Martin. (*Télévision*, no. 51, pp. 45–51; February, 1955.) Illustrated description of a television receiver using a classical objective system for projecting pictures of linear dimensions variable from about 25 cm to 3 m. The equipment is no larger than a standard receiver with 36-cm screen.

621.397.621.2:535.623 2454

Convergence in the CBS-Colortron "205"—J. Giuffrida. (*Radio-Electronic Eng.*, vol. 24, pp. 12–13, 27; February, 1955.) Accurate superposition of the three beams in the center of the screen of this color-television tube is achieved by using radial-deflection correcting coils or permanent magnets on all three guns and a lateral-deflection adjustment on one of them. Convergence of the beams when deflected requires further magnetic fields which vary as complex circular functions of the scanning angle. Circuit arrangements for deriving the required magnetic-field waveforms from the deflection system are indicated.

621.397.7:628.9 2455
Some Problems in Television Lighting—W. C. Pafford. (*Wireless World*, vol. 61, pp. 288–290; June, 1955.) Discussion of scene illumination and camera arrangements.

621.397.74+621.396.97 2456
Present State of U.S.W. Sound and Television Broadcasting Coverage [in Germany]—R. Gressmann. (*Funk-Technik*, vol. 10, pp. 31–34; January, 1955.)

TRANSMISSION

621.376.223 2457
Unbalance Effects in Modulators—D. G. Tucker. (*Jour. Brit. IRE*, vol. 15, pp. 199–207; April, 1955.) Rectifier modulators of (a) shunt ("Cowan") and (b) ring types are considered. With type (a) only one balance control is needed and only one unbalance component can in general be brought to a real minimum by a particular adjustment. With type (b) two independent balance controls can be provided and two unbalance components can be simultaneously brought to a minimum. The effect of the signal voltage on the magnitude of unbalance components is zero in a modulator which is perfectly balanced in the absence of signal, and is noticeable only near the overload point even when initial unbalance exists.

621.396.61:621.3.018.78 2458
The Problem of Distortion in Anode-Voltage-Modulated Transmitters—W. T. Runge. (*Telefunken Ztg*, vol. 27, p. 254; December, 1954.) Correction to paper abstracted in 262 of 1953.

621.396.61:621.375.2 2459
Progress in the Construction of Modulation Amplifiers for Anode-Modulated Broadcasting Transmitters—H. Müller. (*Telefunken Ztg*, vol. 27, pp. 204–210; December, 1954.) Development work was aimed at reducing the amount of equipment and the power dissipation. The essential features of the present circuit arrangement are (a) operation of the driver valve with grid current, (b) dc negative feedback on the driver, and (c) a substantial reduction in the anode currents of the output valves with no signal applied. Because of the higher efficiency resulting from grid-current operation, smaller tubes with lower anode voltage can be used. Measurements on amplifiers for 20-kw and 150-kw transmitters confirmed the saving in equipment and power dissipation.

TUBES AND THERMIONICS

621.314.632:546.289 2460
The Barrier Height of Point-Contact Germanium Diodes inferred from Measurements of the Voltage Dependence of Capacitance—F. F. Roberts and J. R. Tillman. (*Proc. Phys. Soc.*, vol. 68, pp. 113–115; February 1, 1955.) Results of measurements on several types of diode are presented graphically, using a log C/log $(V+\phi)$ presentation so as to give equal prominence to large and small values of the capacitance C and reverse voltage V, the constant ϕ being chosen for each unit so that the curves are nearly straight lines. The value of n, the reciprocal of the slope, is about 2 for most types and about 3 for some others; the interpretation of these values is discussed in terms of the distribution of impurities. It is inferred that for the cases where $n\sim 2$ the barrier height for electrons approaching the semiconductor from the metal is less than or barely equal to half the forbidden energy gap for Ge. Attention is drawn to the discrepancy between this low value of barrier height and the much greater values to be inferred from measurements of injection ratio [e.g. 2536 of 1954 (Swanson)].

621.314.632:546.289:621.317.3 2461
Recovery Time Measurements on Point-Contact Germanium Diodes—Firle, McMahon and Roach. (*See* 2356.)

621.314.7+537.311.33 2462
Semiconductors and the Transistor—E. W. Herold. (*Jour. Frank. Inst.*, vol. 259, pp. 87–106; February, 1955.) A general survey with 33 references.

621.314.7 2463
Some Aspects of the Design of Power Transistors—N. H. Fletcher. (PROC. IRE, vol. 43, pp. 551–559; May, 1955.) Design of alloyed-junction transistors for af operation with output >1 w is considered. Special electrical problems are involved since the injected carrier density cannot be considered as a linear perturbation. Mechanical design problems relate mainly to cooling and shock resistance. The effect of reduction of emitter bias due to transverse current flow in the base region is examined. Transistors with class-A output ratings >50 w have been obtained.

621.314.7.002.2 2464
The Transistor: Part 4—the making of Transistors—H. Yemm and J. L. Carasso. (*P.O. Elec. Eng. Jour.*, vol. 47, Part 4, pp. 217–221: January, 1955.) Part 3: 1070 of May (Speight and Carasso).

621.314.7.012.8 2465
Theory of Equivalent Circuits for Junction Transistors—Oertel. (*See* 2216.)

621.314.7.012.8 2466
The Frequency Dependence of [junction-] Transistor Quadripole Parameters—Kettel and Meyer-Brötz. (*See* 2217.)

621.383.2 2467
Energy Distributions of Photoelectrons from Au and Ge in the Far Ultraviolet—W. C. Walker and G. L. Weissler. (*Phys. Rev.*, vol. 97, pp. 1178–1179; February 15, 1955.)

621.383.2 2468
Evolution and Conservation of the Photoelectric Effect of an Sb-Cs Layer deposited in Vacuum—A. Lallemand and M. Duchesne. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1329–1331; March 21, 1955.] Experiments are described in which a sealed tube containing an Sb-Cs photocathode is introduced into an evacuated enclosure. After evaporating Ba onto the enclosure wall, the inner tube is broken and the time variation of the photoelectric emission is observed. Photosensitivity is retained longer by those cathodes which are richer in Cs. When the cathode is placed on a support cooled to about -55 degrees C. the photosensitivity is not appreciably reduced after several hours, while the dark emission is considerably reduced.

621.383.27 2469
Theory of Operation of the Final Stage of a Photomultiplier—P. Leuba. (*Jour. Phys. Radium*, vol. 16, pp. 161–162; February, 1955.) Under specified operating conditions, the voltage variation on the collector electrode is expressed approximately by $du/dt+u/RC=q(t)/C$, where RC is the time constant of the equivalent capacitor, and $q(t)$ the charge transferred from the penultimate electrode to the collector. When $RC/T\gg 1$, T being the period of charge transfer, operation is linear. When $RC/T<100$, experiment shows that $q(t)$ is approximately an inverse exponential function.

621.383.4/.5:546.289 2470
Germanium Junction Photodiodes—Zh. I. Alferov, B. M. Konovalenko, S. M. Ryvkin, V. M. Tuchkevich and A. I. Uvarov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 11–17; January, 1955.) In the photocell proposed by Shive (2825 of 1953) the direction of the beam of light is parallel to the plane of the n-p junction. In the cell used in the present investigation the light beam is perpendicular to the plane of the junction and passes through a thin layer of n- or p-type Ge. Experiments were carried out to determine the voltage/current characteristics, dependence of current on intensity of illumination, distribution of sensitivity over the surface, spectral distribution of sensitivity, inertia and temperature effect.

621.383.4/.5:546.289 2471
Sensitivity of Germanium Photodiodes to X Rays—B. M. Konovalenko, S. M. Ryvkin and V. M. Tuchkevich. (*Zh. Tekh. Fiz.*, vol. 25, pp. 18–20; January, 1955.) Measurements indicate that Ge photodiodes are much more sensitive to X rays than silver sulphide or selenium photocells. They also possess the advantage of a linear current/irradiation characteristic.

621.383.4/.5:546.289 2472
Mechanism of Operation of Germanium Photodiodes—S. M. Ryvkin. (*Zh. Tekh. Fiz.*, vol. 25, pp. 21–28; January, 1955.) Theory of the operation of an n-p Ge diode as a photocell is given; the fundamental relations are established between the photo-emf, short-circuit photocurrent and saturation current for the case of operation as a barrier-layer photocell. A general equation suitable for any operating conditions is also derived. Results are given of experiments made to verify the relations obtained, and the efficiency of the diodes used as barrier-layer photocells is discussed.

621.385:621.317.7 2473
Circuit for the Determination of Contact Potentials and Electron Temperatures from Retarding Field Characteristics—Friedman and Heynick. (*See* 2369.)

621.385:621.396.822 2474
Fluctuations of Cathode Emission in Electron Valves—A. Blanc-Lapierre, G. Goudet and P. Lapostolle. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 1409–1411; March 28, 1955.] Noise due to fluctuations of electron velocity and of current density is discussed. Separate consideration is given to the cases where (a) electrons are emitted at regular intervals, and (b) electrons are emitted at irregular intervals. In case (a) there is no noise contribution from current fluctuation; expressions are given for the current-fluctuation contribution in case (b) and for the velocity-fluctuation contribution, which is twice as great in case (b) as in case (a). Values derived for the anode-current fluctuations of a space-charge-controlled diode, using the case (a) formula, are in good agreement with those obtained by direct calculation and in fairly good agreement with values obtained experimentally, indicating that the electron emission is regulated by the virtual cathode.

621.385.029.6 2475
Space-Charge Waves in Ion-Free Electron Beams—J. Labus and K. Pöschl. (*Arch. elekt. Übertragung*, vol. 9, pp. 39–46; January, 1955.) In a previous analysis by Labus (2182 of 1953) the rf variations of the beam boundary were neglected; on taking these into account, the theoretical results are in satisfactory agreement with measurements made at the Massachusetts Institute of Technology. See also 2807 of 1954.

621.385.029.6 2476
The Design of Travelling-Wave Output Valves for Microwave Relay Stations—W. Klein. (*Arch. elekt. Übertragung*, vol. 9, pp. 55–62; February, 1955.) The operation of helix-type valves capable of giving outputs of the order of 5 w is discussed on the basis of the relation between gain and power level. To obtain the required output it is necessary to use electron guns providing strong beam con-

centration, or alternatively to use cathodes giving high current density; the requirements become more difficult to satisfy as the wavelength decreases. With high-gain valves very good matching is required between the localized attenuation section and the rest of the helix. Typical values of efficiency are 10–15 per cent. See also 1526 of June (Klein and Friz).

621.385.029.6:621.317.755 2477

Velocity Spectrography of Electron Dynamics in a Travelling Field—Purl and Von-Foerster. (*See* 2379.)

621.385.032.2:537.533 2478

New Approach to the Design of Electron Guns for Cylindrical Beams with High Space Charge—M. Müller. (*Arch. elekt. Übertragung.* vol. 9, pp. 20–28; January, 1955.) Discrepancies between the predicted and observed characteristics of guns with spherically curved electrodes (Pierce type) are studied. The influence of the anode aperture is evaluated by electrolyte-trough measurements, using specially shaped anode and control electrodes. Design details are given for simple electrode shapes and arrangements, and an indication is given of the corrections for high perveance values. Effects due to thermal velocity distribution are discussed.

621.385.032.21 2479

The Platinum-Cored Oxide-Cathode Repeater Valve—G. H. Metson. (*P.O. Elec. Eng. Jour.*, vol. 47, part 4, pp. 208–211; January, 1955.) The reactions of cathodes with pure platinum and nickel cores to oxygen attack are compared. Complete recovery from heavy oxygen contamination occurs in the case of platinum, whereas the nickel-cored cathode is completely destroyed.

621.385.032.213 2480

Deviation from the Boguslavski-Langmuir Law when a Tungsten Cathode of a Thermionic Valve is Heated by a High-Density Current Pulse—S. V. Lebedev. (*Zh. Eksp. Teor. Fiz.*, vol. 27, pp. 487–500; October, 1954.) Experiments were made to find the cause of the abnormal increase in the anode current when a tube filament is heated by a high-density current pulse. A detailed report is presented including photographs of oscillograms recorded. The main conclusion reached is that the current increase is due to a change in the state of the tungsten, and is not related to the neutralization of the space charge by ions.

621.385.032.216 2481

Electron-Optical Study of Non-stationary Emission from an Oxide Cathode in Vacuum and in a Gas—I. N. Prilezhaeva, V. V. Livshits and G. V. Spivak. (*Zh. Tekh. Fiz.*, vol. 25, pp. 97–107; January, 1955.) The emission from a pulsed cathode was studied. Electron-optical images were obtained showing the disturbances caused by sparking and cathode poisoning. Oscillograms of discharge currents corresponding to these images are also given.

621.385.032.216 2482

Growth of the Barium Orthosilicate Interface of Oxide-Coated Cathodes—M. G. Harwood and N. Fry. (*Brit. Jour. Appl. Phys.*, vol. 6, pp. 62–64; February, 1955.) The core/coating interface of oxide cathodes with Ni-alloy cores containing (a) 0.048 per cent and (b) 0.17 per cent Si was examined. In addition to barium orthosilicate a layer of nearly pure strontium oxide was found. Results of an investigation of the barium orthosilicate layer during heating at 765 degrees C. for up to 2,000 hours indicate that the growth continues until all the available Si is used up, in case (a), but a limitation of the growth is apparent in case (b), possibly due to diffusion of Si.

621.385.832 2483

Optical Distortion of Magnetic Deflecting Coils—E. Cambi. (*Alta Frequenza*, vol. 23, pp. 292–334; December, 1954.) First-order approximations are made for three distinct causes of distortion considered separately. These are (a) finite length and axial nonuniformity of the field, (b) curvature of the screen, and (c) transverse nonuniformity of the field. (a) is substantially independent of the direction of deflection, and may be partly compensated by (b) provided the screen is rotationally symmetric and concave. (c) is considered in terms of percentage of field harmonics present, i.e. in terms of design parameters of the coil. Conditions under which the geometric and third-harmonic distortion due to (c) can cancel out in an axial direction are established. All formulas are expressed in terms of design parameters of a cr tube.

621.385.832 2484

A Revolutionary Television Tube—(*Radio and Telev. News*, vol. 53, p. 44; April, 1955.) Brief description of a very-short-length picture tube comprising a phosphor screen sandwiched between glass plates. An electron beam is directed along an edge, adjacent to a row of deflection plates; by applying a control voltage to a selected deflection plate the beam is bent at right angles, remaining parallel to the plane of the screen. A second set of deflection plates is used to bend the beam so as to strike the screen normally.

621.385.832:535.37 2485

Reduction of Cathodoluminescence of ZnS:Ag by Irradiation with 16-kV Electrons—K. H. J. Rottgardt and W. Berthold. (*Naturwiss.*, vol. 42, p. 67; February, 1955.) Experiments were made using ordinary 43-cm television tubes with Al-coated screens. Comparison of the results with those obtained previously for bombardment with 4-kv electrons [3420 of 1954 (Rottgardt)] confirms that the reduction of luminescence depends only on the number of incident electrons and is independent of electron energy over the range 2–16 kv.

621.385.832:681.142 2486

A Beam-Deflection Valve for use in Digital Computing Circuits—M. W. Allen. (*Proc. IEE*, Part C, vol. 102, pp. 57–61; March, 1955. Digest, *ibid.*, Part II, vol. 101, pp. 682–684; December, 1954.) The properties required in a universal computing element are discussed in terms of the analogy with the nervous system. A description is given of a ribbon-beam-deflection tube consisting of two interconnected elements each having the desired properties; this enables binary addition and other operations of interest to be performed in a single tube with a 12-pin base.

621.387 2487

Restoration of Control in Ionic Apparatus—V. I. Drozdov and A. F. Smirnov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 85–96; January, 1955.) The time necessary for the restoration of control in devices such as thyratrons, mercury rectifiers etc. after the cessation of the current is considered; various definitions are discussed. An elementary theory of the restoration process is proposed and two experimental methods are described for determining the "electric-strength"/time build-up characteristic which indicates the ability to block the anode voltage. Some experimental results are given.

621.387.032.216 2488

Peak Current of Oxide Cathodes in Arc Discharges—H. J. Vogt. (*Elektrotech. Z.*, *Edn A*, vol. 76, pp. 192–195; March 1, 1955.) An oscillographic pulse-method for determining the permissible peak current is described. At this value of current, when the emission from the cathode just becomes nonuniform, both the anode voltage and the emission current cro traces show sharp "spikes." This condition was also observed spectroscopically. A circuit diagram of the experimental setup is given.

MISCELLANEOUS

061.4:[621.317.7+621.38 2489

Physical Society's Exhibition [1955]—(*Wireless World*, vol. 61, pp. 271–278; June, 1955. Correction, *ibid*, vol. 61, p. 324; July, 1955.) An illustrated review of exhibits, including instruments and tubes shown also at the RECMF exhibition (2490 below). *See* also *Wireless Eng.*, vol. 32, pp. 169–172; June, 1955.

061.4:621.396.6 2490

Components Exhibition—(*Wireless World*, vol. 61, pp. 258–264; June, 1955.) Detailed review of components and accessories at the RECMF exhibition held in London, April, 1955. *See* also 2489 above.

621.19:621.37/.38].004.4 2491

Mould Growth in Electronic Apparatus—E. Ganz and O. Walchli. (*Bull. schweiz. elektrotech. Ver.*, vol. 46, pp. 233–239; March 19, 1955. In French.) The IEC-recommended methods for testing equipment and components for resistance to mold growths are discussed and modified test specifications are proposed based on experimental results. Spraying or painting with a suitable varnish should give adequate protection.

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(Continued on page 100A)

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(Continued from page 98A)

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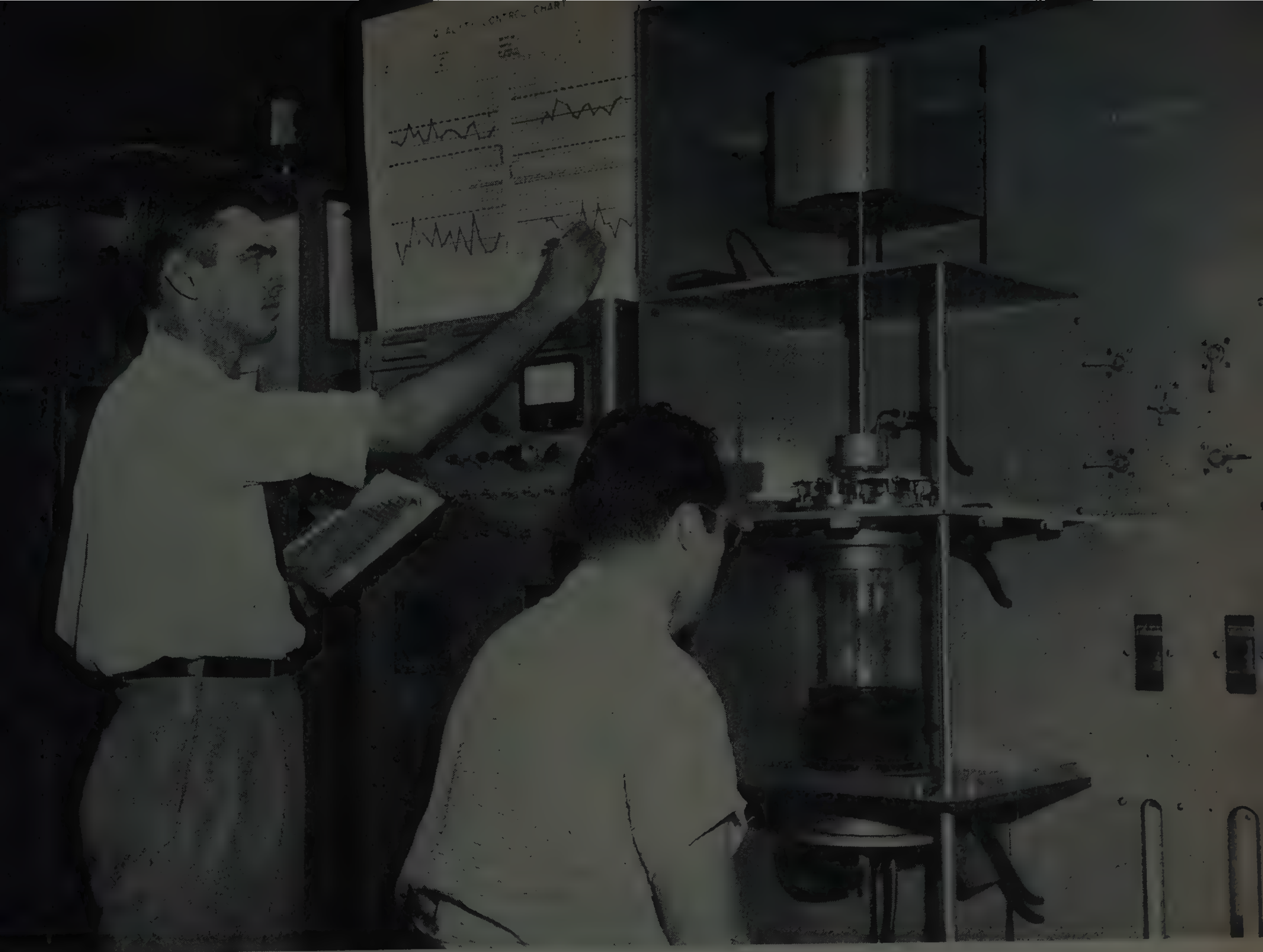
(*Continued on page 104A*)



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(*Continued on page 106A*)

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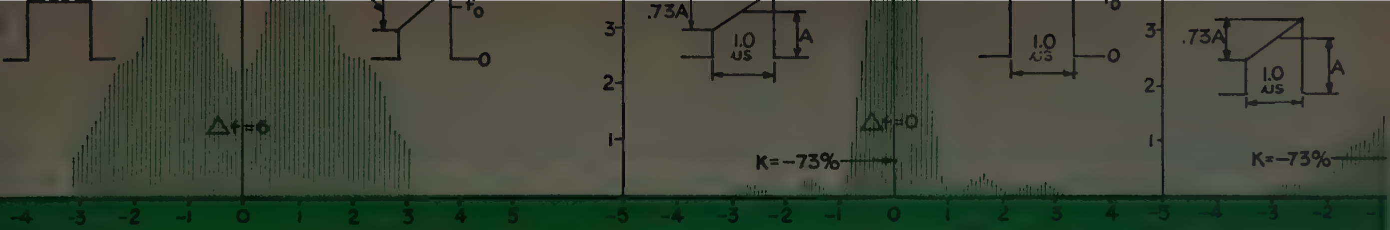
(Continued on page 108A)



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SPECIFICATIONS

DIMENSIONS
25-7/16″ high by 20⅛″ wide by 19⅛″ deep.

WEIGHT
150 pounds

PRESENTATION
5CPIA 5″ cathode-ray tube (other persistances available).

SENSITIVITY
At signal to noise ratio 2:1, and spectrum width 25 megacycles:
— 75 dbm at 10 mc to
— 50 dbm at 16,000 mc

RANGE
10 megacycles to 16,000 megacycles calibrated. Usable 1 to 34,000 megacycles.

ACCURACY
Dial accuracy ±1.0% at the operating frequency of the local oscillator.

SPECTRUM WIDTH
0.5 to 25 megacycles

RESOLUTION
10 kilocycles

TEMPERATURE RANGE
Operating — 40 to + 130° F

HUMIDITY
90% RH.

SHOCK
(Non-operating in transit case.) One 12G impact, 10 mlsec duration on each face. One 37G impact, 10 mlsec duration on

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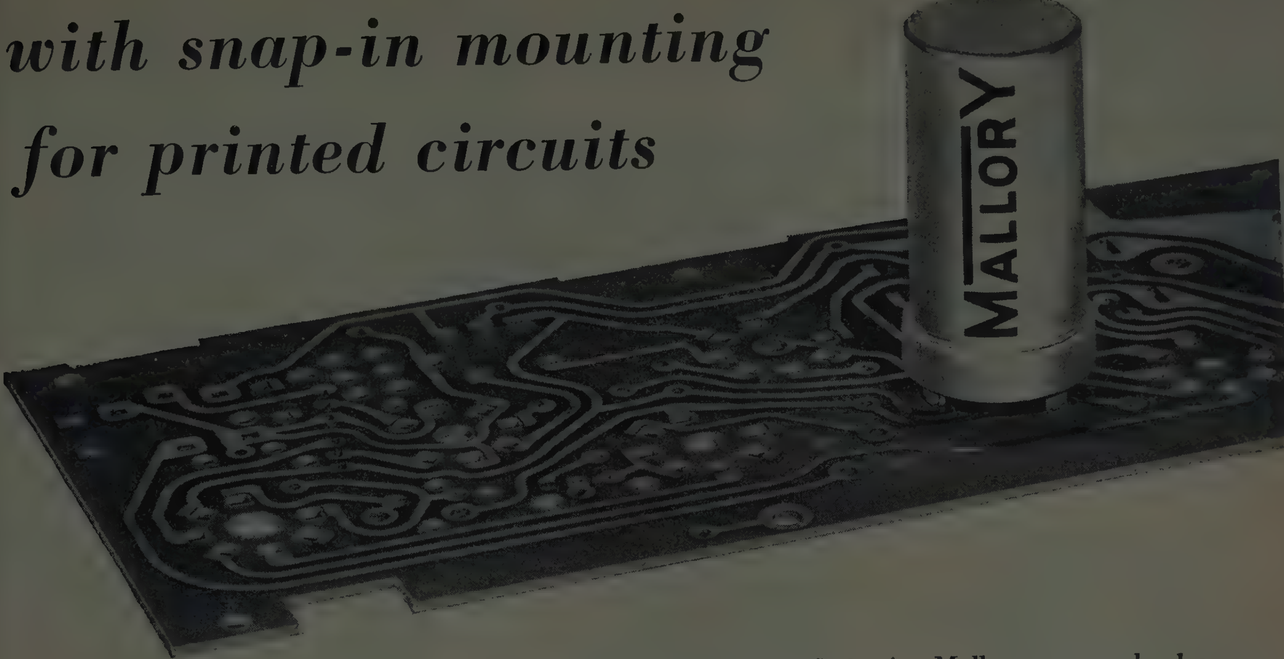
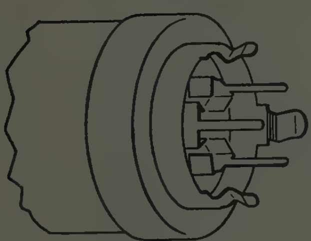
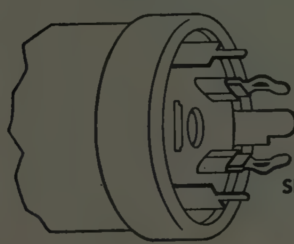
New MALLORY FP Capacitors

with snap-in mounting for printed circuits





SNAP-IN MOUNTING TABS



SNAP-IN TERMINALS

If you are using printed circuits, Mallory can supply electrolytic capacitors with the terminal construction you need. During nearly two years of developing and manufacturing capacitors especially for printed circuit use, Mallory has created a diversified group of designs that cover most applications.

The latest additions to the line of FP Capacitors for printed circuits are designed for snap-in mounting. Just push the capacitor into its slots in the circuit panel, and spring-formed tabs hold it in place, ready for soldering.

You have a choice of either snap-in mounting tabs or snap-in terminals. In addition, you can select models with straight tabs and terminals. All are available in six-slot or eight-slot terminal configurations.

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Circuits can be printed on both sides. Shoulders on the mounting tabs hold the capacitor case clear of the printed sheet. Clearance ranges up to .137″.

Positive soldering. Possibility of aluminum contamination is eliminated because the connections from the foil stop well short of the solder area.

Added to these time-saving design features are the superior electrical characteristics and long life at high temperatures which have made Mallory FP Capacitors the standard of performance throughout the industry. Write or call us today for technical data, and for an analysis of your circuit requirements by a Mallory capacitor engineer.

MALLORY
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for Printed Circuits

In addition to the FP line of capacitors, Mallory produces a special series of metal tubular electrolytics for printed circuits. One terminal is a bare wire, and the other a flat tab for orientation. Write or call for technical data and available ratings.

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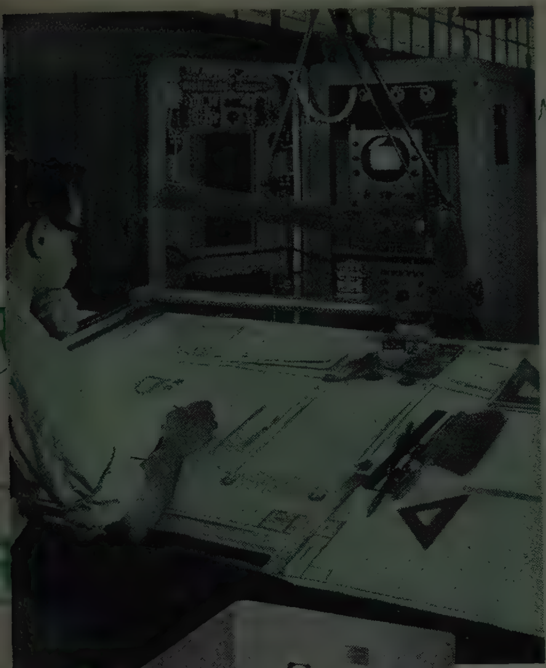

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Z822A	0.2	20	10	3,330	.018	10/1200-1800	3/440	6x3x4	3330	SAL 39 (GL-6625)
Z823A	0.042	12.7	3.3	3,330	.064	coded/300	3/440	6x4x4	3200	SAL 39
Z824A	0.625	25	25	1,000	.001 at 0.5/2000	1.0/1000 2.0/250 5.0/100	1/190-250	6x2x3	2500	QK-349 at reduced power
Z825A	1.6	40	40	1,000	.001 at 0.5/2000	1.0/1000 2.5/400	1/190-250	6x2x3	2500	WE-5780

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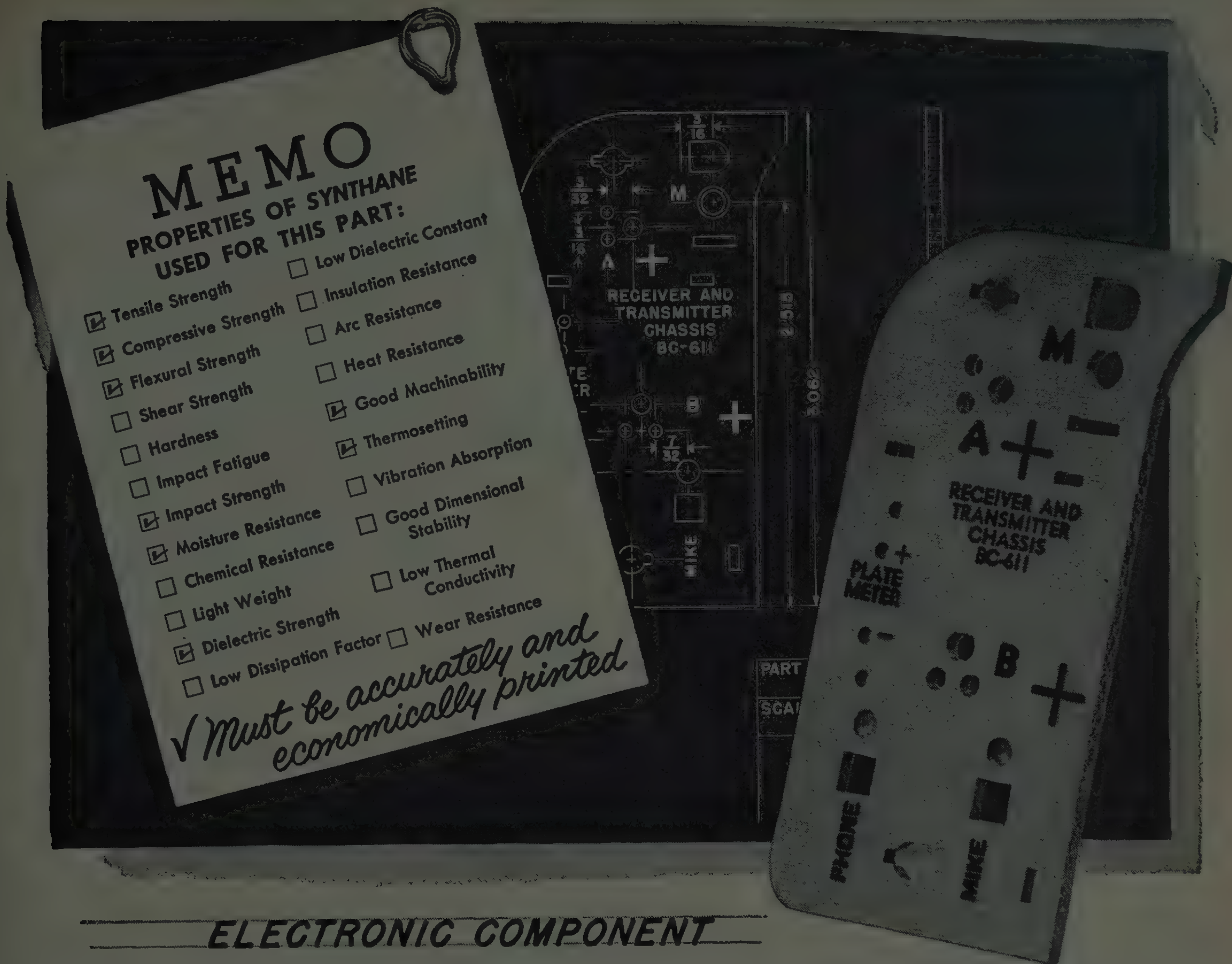

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City__________ Zone_____ State_____

SYNTHANE
LAMINATED **S** PLASTICS

SYNTHANE CORPORATION • OAKS, PENNSYLVANIA

IRE Membership

(Continued from page 114A)

Spergel, J., 412 Third Ave., Asbury Park, N. J.
Spieker, L. J., 6000 Lemmon Ave., Dallas 9, Tex.
Spielberg, A. M., 267 Crystal Terr., Haddonfield, N. J.
Spinks, A. W., 4216 Knowles Ave., Kensington, Md.
Spinner, L. N., 3867 Maywood Court, Cheviot 11, Ohio
Spoor, T. A., Philips Mexicana S. A. Calle Norte 44, No. 3612, Col Xochimanca-Atzcapotzalco, Mexico 16, D. F. Mexico
Sprague, R. M., 76 Annawan Rd., Waban 68, Mass.
Sprague, V. G., 31 Brunswick Rd., Montclair, N. J.
Sproul, S., R.F.D. 1, Cato, N. Y.
Sprunger, R. W., 809 S. Grove Ave., Oak Park, Ill.
Squibb, W. F., 3910 Alicia Dr., San Diego 7, Calif.
Squitieri, A., 70 Crest Rd., Merrick, L. I., N. Y.
Sreb, J. H., 203 Elmira St., S.W., Washington 24, D. C.
Stahl, M. D., Hoover Co., North Canton, Ohio
Stahmann, J. R., 1827 Carl St., St. Paul 8, Minn.
Stambaugh, D. W., 1528—28 St., Columbus, Ind.
Stanton, G. V., 1257 Punta Way, Monterey Park, Calif.
Stanton, R. S., 10747 Rochester Ave., Los Angeles 24, Calif.
Stanton, S. W., Halsey Rd., East, Box 383, Parsippany, N. J.
Stanwick, C. A., 131 Rynda Rd., South Orange, N. J.
Starek, R. A., 412 Highview Ave., Elmhurst, Ill.
Starks, W. H., 7203 Leonard St., Philadelphia 49, Pa.
Starner, C. J., 314 Elm Avenue, Haddonfield, N. J.
Staschover, L., 211-01—75 Ave., Bayside 64, L. I., N. Y.
Stavitski, L. X., 3705—64 St., Woodside, L. I., N. Y.
Steeb, E. C., Jr., 15 Groveland Ave., Buffalo 14, N. Y.
Steele, E. R., 55 Harding Ave., Hatboro, Pa.
Stein, E., 31-07 Garrison Terr., Fairlawn, N. J.
Stein, S., 90 Day St., Clifton, N. J.
Steinberg, W. A., 3756 Libby Lane, Levittown, L. I., N. Y.
Stenhammer, H. T., 55 E. 93 St., New York, N. Y.
Stensgaard, C. P., Jr., 240 S. Michigan Ave., Pasadena 5, Calif.
Stephanz, G. H., 8 Lake Dr., Darien, Conn.
Stephenson, J. O., 127 Dumbarton Rd., Baltimore, Md.
Sterk, A. A., 198 Seminary Ave., Yonkers, N. Y.
Stern, J. L., C.B.S. Engr., Dept., 485 Madison Ave., New York 22, N. Y.
Sternick, L., 3451 Giles Pl., New York 63, N. Y.
Stetson, H. T., 541 Lido Dr., Ft. Lauderdale, Fla.
Stevens, W. J., 253 Governors Ave., Medford 55, Mass.
Stewart, B. L., Insp. of Naval Material, 2300 Eleventh Ave., S.W., Seattle 4, Wash.
Stewart, J. A., 23 Drake, Mount Kisco, N. Y.
Stewart, W. A., 5816 Trojan Ave., San Diego 15, Calif.
Stiles, M. G., 789 Prospect St., Maplewood, N. J.
Stimmel, R. G., 141 Foxridge Dr., Dayton 9, Ohio
Stoll, J. L., 1031 N. Lake St., Madera, Calif.
Stoltz, G. F., 808 S.E., Riverside Dr., Evansville 13, Ind.
Stone, W. R., 21 Horn Lane, Levittown, L. I., N. Y.
Stonevic, V. R., 1342 Argyle, Chicago 40, Ill.

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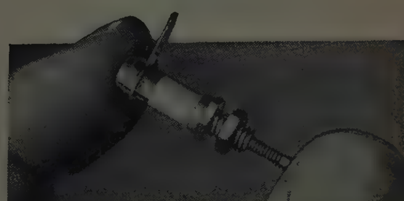

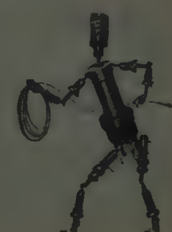

Membership

(Continued from page 116A)

Stormer, W. J., 2800—51 Ave., S.E., Washington 27, D. C.
Stotz, C. C., 180 Pine St., Rockville Centre, L. I., N. Y.
Stouse, L. E., Jr., 4233 E. Second St., Long Beach 3, Calif.
Stout, E. R., 620 Clarendon Ave., Arlington Heights, Ill.
Stovall, H. E., 3909 First Ave., N.E., Seattle, Wash.
Stover, E. B., Jr., 2301 S. Florence Pl., Tulsa 5, Okla.
Strain, C. E., 1298 Niles Ave., N.W., Atlanta, Ga.
Strain, M., 64 Hickory St., Dumont, N. J.
Strand, J., 3051 Geddes Ave., Ann Arbor, Mich.
Straney, K. O., 20 Pearl St., Ext., Beverly, Mass.
Streater, A. L., 120 Sylvan Dr., Broomall, Pa.
Streich, R. J., Box 121, Inverness, Calif.
Streifus, C. A., 2 Rector St., Rm., 930, New York 6, N. Y.
Stribling, J. L., Jr., 2003 Columbus, Ft. Worth, Tex.
Striker, R. G., 58 South Dr., Plandome, L. I., N. Y.
Strobach, F., 106 Myrtle St., Suite 9, Boston 14, Mass.
Stroble, R. R., 4715 Chestnut St., Bethesda 14, Md.
Stroman, C. F., Officers Mail Sect., 814 Opns. Sq., Fairchild Air Force Base, Fairchild, Wash.
Strother, J. F., Hq. AMO, Box 4195, Wright Patterson AFB, Ohio
Strothers, H., 4111 33 St., S., Arlington, Va.
Stroud, S. G., 3501 McBain Ave., Redondo Beach, Calif.
Strull, G., 819 College Ave., Pittsburg 32, Pa.
Stuart, R. M., 270 Ocean Ave., Marblehead, Mass.
Stuber, Dr. R., Schlosslistr, 29, Zurich 44, Switzerland
Stubner, J. W., 2116 Grove St., Glenview, Ill.
Suarez-Flamerich, O., Qta. Mariaemma Ave., Valle Arriba, Caracas, Venezuela
Subramaniam, C. R., Asst. Supt. Development, Tech., Devel., Estab., Sunderwala Camp., Dehra Dun, U.P., India
Sukup, J. P., 4317 W. 99 Pl., Oak Lawn, Ill.
Summers, G. C., 1447 Fleming Ave., Dallas 16, Tex.
Summers, W. P., 1540 Glenwood Rd., Vestal, N. Y.
Sunder, J. E., 320 Duckpond Dr., South, Wantagh, L. I., N. Y.
Sundquist, A., El. Engr., Stenkullavagen 45, Stora Essingen, Stockholm K. Sweden
Sutton, G. L., Box 313 Griffiss AFB, Rome, N. Y.
Sutton, J. G., Jr., 1524 E. Oak St., Phoenix, Ariz.
Svedlow, J. D., 102-1121 Patrick Henry Dr., Falls Church, Va.
Svihel, B. T., 940 Wyoming Ave., Croydon, Pa.
Swain, R. R., 2324 General Bradley N.E., Albuquerque, N. M.
Sweeney, W. R., 32 Wyatt Rd., Garden City, L. I., N. Y.
Swindall, P. M., P.O. Box 682, Huntsville, Ala.
Swire, B. E., 78 Arabella St., Longueville, N.S.W., Australia
Synnamon, J. H., 255 Eberly Ave., Bowling Green, Ohio
Taber, J. E., 2804 W. 163 St., Gardena, Calif.
Tanck, H., 197 Wicks Lane, Malverne, L. I., N. Y.
Tanigawa, G., Telechron Inc., Gen. Elec. Co., Ashland, Mass.
Tanner, W. E., Bell Aircraft Corp., Buffalo 5, N. Y.

(Continued on page 120A)

Membership

(Continued from page 118A)

Taris, C. M., 11 Adams Ave., Cranford, N. J.
Tasker, R., 7000 Lasaine Ave., Van Nuys, Calif.
Taylor, E. R., Bell Tel. Labs., Murray Hill, N. J.
Taylor, E. S., Hedge Apple Lane, Wilmington, Del., RFD 1
Taylor, H. M., Jr., 411 W. Holmes, Huntsville, Ala.
Taylor, K. A., Bell Tel., 416 Seventh Ave., Pittsburgh 19, Pa.
Taylor, Malcolm, 6705 Sherwood Rd., Baltimore 12, Md.
Taylor, P. H., 224 Washington Ave., Chatham, N. J.
Temple, D. I., 600 Pelham Rd., Bldg. B., New Rochelle, N. Y.
Tenenbaum, F., 55 Plymouth Rd., Great Neck, L. I., N. Y.
Tennant, M. J., 2159 Five Mile Line Rd., Penfield, N. Y.
Teske, J. A., 115 Sheffield St., Silver Spring, Md.
Teslik, J. A., 3854 W. Van Buren, Chicago 14, Ill.
Thaxton, Herbert A., 4029 S. Acoma, Englewood, Colo.
Theobald, D. J., 5450 Linda Rosa, La Jolla, Calif.
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Thille, R. W., 1 Wood St., Apt. 7, San Francisco 18, Calif.
Thomas, E. R., 245 Maple St., Englewood, N. J.
Thomas, M. E., 575 'F' St., Idaho Falls, Idaho
Thomas, R. R., Jr., Radio Sta. W.O.A.Y., Oak Hill, W. Va.
Thomas, T. R., 14834 Coyle Ave., Detroit 27, Mich.
Thomson, B., Jr., 814½ N. Madison, Pasadena 6, Calif.
Thompson, D. C., 30 Cleveland Rd., Waltham 54, Mass.
Thompson, G. L., 7104 Flight Ave., Los Angeles 45, Calif.
Thompson, J. F., 27 N. Fernwood Dr., Rockledge, Fla.
Thompson, J. E., 1440 W. 47 St., Chicago 9, Ill.
Thomson, E. C., 55 Livingston Rd., Wellesley 82, Mass.
Thurensen, R., 11420 Bolas St., Los Angeles 49, Calif.
Thrasher, P. M., Jr., 1612 Eastman Dr., Falls Church, Va.
Thunen, G. W., Box 182, Orinda, Calif.
Thurston, E. G., 2433 N. Belvoir Blvd., Cleveland 21, Ohio
Tibbetts, G. C., Colcord Ave., Camden, Maine
Tillotson, J. G., 2532 Place De Monceau, Apt. 5, Sillery P.Q., Canada
Tilton, D. S., 27 Pennacook St., Manchester, N. H.
Timineri, A. A., 3518 Lenox Dr., Dayton 9, Ohio
Tingley, E. M., Jr., 383 Elm St., Glen Ellyn, Ill.
Tipton, C. D., Rt. 3, Box 269½, Austin, Texas
Tittle, H. C., 81 Rose Hill Way, Waltham 54, Mass.
Tobin, M. W., 58-02—251 St., Little Neck, L. I., N. Y.
Tobler, C. R., 20 Sutton Pl., Verona, N. J.
Tongue, B. H., 526-536 N. Ave., Westfield, N. J.
Torran, J. T., 88-11—34 Ave., Jackson Heights, L. I., N. Y.
Tourshou, S., 1716 Spring Garden St., Philadelphia 30, Pa.
Trachtenberg, R., 401 Graisbury Ave., Haddonfield, N. J.
Trapp, R. R., 320 W. Franklin St., Taylorville, Ill.

(Continued on page 121A)

(Continued from page 120A)

Trinter, V. E., 2808 Echodale Ave., Baltimore 14, Md.
Trotter, B. E., 3110 First St., N., Arlington, Va.
Truscott, D. N., 58 Oxford Rd., Birmingham 13, England
Tuckerman, L. S., 317 E. 87 St., New York 28, N. Y.
Tunnicliffe, W. W., 11 Orient St., Winchester, Mass.
Turner, A. H., 202 Burrwood Ave., Collingswood 7, N. J.
Turner, A., 160 Gaylor Rd., Scarsdale, N. Y.
Tuttle, L., 1131 Ashburton St., Baltimore 16, Md.
Tutty, J. E., 56 Jefferson St., Belleville, N. J.
Tykulsky, A., 551 Wales Ave., New York 55, N. Y.
Ulmer, R. M., 2207 Campus Rd., Beachwood 21, Ohio
Umbach, L., Jr., 21 Olive St., Youngstown 1, Ohio
Urey, G. M., 934 N. Orlando Ave., Hollywood 46, Calif.
Vacca, L. N., 7 Meadowbrook Lane, Vestal, N. Y.
Valenta, M. F., 14 Ben Oaks Dr., W., Millersville, Md.
Vallarino, A. R., E. Saddle River Rd., Allendale, N. J.
Vallarino, A. D., Casilla 1753, Santiago De Chile
Vanacore, T., 151 Tulane Rd., Kenmore 17, N. Y.
Vance, D. H., RD 1, Crooked Lane, Moorestown, N. J.
Van Doeren, C. A., Peace Haven Rd., Winston-Salem, N. Car.
Van Every, A., 1017 S. California, Albuquerque, N. Mex.
Van Meter, J. L., 4298 Highview Ave., Baltimore 29, Md.
Van Name, J. M., 1506 E. Walnut Lane, Philadelphia, Pa.
Vasilevskis, H. S., Tioga & C Sts., Philadelphia 34, Pa.
Vaughan, V., 3836 Pukalani Pl., Honolulu, T. H.
Veazie, E. A., Bell Tel. Labs., Murray Hill, N. J.
Vehling, R. H., 60 Hudson St., New York 13, N. Y.
Vendeland, R. N., 610 Fourth St., Riverton, N. J.
Verkley, Bernard M., 3768 Hubble Rd., Cincinnati 24, Ohio
Vetter, William F., 43 Sprucewood Dr., Levittown, L. I., N. Y.
Viales, Lionel O., 2332 Cedar St., Berkeley 8, Calif.
Vilas, Edward T., Box 491, Silver Spring, Md.
Vogel, H. O., U.S.S. Rehoboth, c/o F.P.O., New York, N. Y.
Vogel, W. H., Jr., 59 Crane St., Caldwell, N. J.
Von Dohlen, H. W., 1701 Preston Rd., Alexandria, Va.
Vore, M. P., c/o Westinghouse Elec. Corp., Baltimore 3, Md.
Vulliet, P. O., Rt. 2, Box 870, Bremerton, Wash.
Waer, R. R., 49 Daily St., Nutley, N. J.
Wagner, F. C., 16219 Tonkaway Rd., Wayzata, Minn.
Walden, G. R., 3804 Fulton St., N.W., Washington 7, D. C.
Wales, R. O., 301 Vierling Dr., Silver Spring, Md.
Walkup, L. A., 4904—14 Ave., S., Minneapolis 17, Minn.
Wall, V. W., 10058 Lesterford Ave., Downey, Calif.
Wallace, Bert E., Jr., 7509 Far Hills Dr., Towson 4, Md.
Wallace, Don C., Bendix Bldg., 1206 Maple Ave., Los Angeles 15, Calif.

(Continued on page 122A)

Membership

(Continued from page 121A)

Wallace, Robert M., 2921 Cherrywood, Dallas 19, Tex.
Wallenstein, J. P., 89 Shephard Ave., Newark 8, N. J.
Walsh, J. V., 501 E. 234 St., Apt. 4E, New York 66, N. Y.
Walter, G. W., 4425 Main St., Kansas City, Mo.
Walters, Andrew W., National Bur. of Standards, Corona, Calif.
Wamboldt, H., 601 Meyer Ave., Dayton 3, Ohio
Wang, C. S., 1205 Eudora St., Denver 20, Colo.
Ward, D. M., 2926 Balboa Ave., San Diego 9, Calif.
Warnick, A., 12752 Memorial, Detroit 27, Mich.
Warren, M., Jr., Box 465, Alcoa, Tenn.
Washburn, C. A., Andria Lane, Thornwood, N. Y.
Wasserman, J., 760 Bronx River Rd., Bronxville, N. Y.
Wasson, I., 8 Scooter Lane, Hicksville, L. I., N. Y.
Waterman, P., Naval Research Lab., Anacostia Sta., Washington 25, D. C.
Watson, R. H., 3741 Narragansett, San Diego 7, Calif.
Watterberg, J., 1227 Garcia, N.E., Albuquerque, N. Mex.
Wayer, C. F., Jr., Star Rt., Box 2805, Spenard, Alaska
Weaver, A., 463 W. St., New York 14, N. Y.
Webb, R. H., c/o Sperry Products Co., Shelter Rock Rd., Danbury, Conn.
Weber, J. W., 8819½ Belford Ave., Los Angeles 45, Calif.
Weber, M. E., 1933 Santa Rose, Houston 23, Tex.
Weeks, G. E., Bell Tel. Labs., Inc., Whippany, N. J.
Wehner, G. L., Olive Rd., Box 385, RR 10, Dayton, Ohio
Weidemann, H. K., 26 Chesapeake St., N.W., Apt. 5, Washington 24, D. C.
Weidknecht, C. J., Jr., 1611 Roumfort Rd., Philadelphia 19, Pa.
Weiler, G., 1429 E. Monroe, South Bend, Ind.
Weill, A. E., 1229 Blanchan Ave., La Grange Park, Ill.
Weinberg, D., 7087 E. Glenmeadow Lane, Cincinnati 37, Ohio
Weingarten, J., 114 Farnham St., Belmont 78, Mass.
Weinstein, A., Box 478, RFD 5, Alexandria, Va.
Weinstein, G., 5251 Pullman Ave., Seattle 58, Wash.
Weinstein, M. S., 1408 Langley Way, Hyattsville, Md.
Weise, D., WTTW-TV, 38 S. Dearborn St., Chicago 3, Ill.
Weiss, L. M., 857 Crotona Pk., N.,, Bronx 60, N. Y.
Welan, I., Hillside Ave., Spring Valley, N. Y.
Welch, H. E., 3301 Kensington Way, Stockton, Calif.
Welker, John J., 2553 N. Mason Ave., Chicago, Ill.
Wellman, B., Box 88, Lincoln, Mass.
Wells, W. M., 3222 Ibsen St., San Diego 6, Calif.
Wells, W. S., 1207 Ryan Ave., St. Paul 13, Minn.
Wendt, R. H., 624 Garrett Pl., Evanston, Ill.
Wengryn, M., 248-11—87 Dr., Bellerose, L. I., N. Y.
Werman, H. M., 1014 E. 24 St., Brooklyn 10, N. Y.
West, J. L., 7 Edwards St., Binghamton, N. Y.
Westbom, A. C., Jr., 16 Circuit Rd., Westwood, Mass.
Westenhaver, J. L., Hughes Aircraft Co., Florence & Teale Sts., Culver City, Calif.

(Continued on page 124A)

Membership

(Continued from page 122A)

Westman, H. P., Jr., 55 Arleigh Dr., Albertson, L. I., N. Y.
Whannel, R. L., R.F.D. 1, Oakdale, Conn.
Wheat, Wilson M., Box 266, MPO, Springfield, Mo.
Wheeler, Randolph B., 130 La Rue Dr., Huntington, L. I., N. Y.
White, A. D., Woodridge Rd., Wayland, Mass.
White, E. A., 1204 Delta Blvd., Ft. Wayne 3, Ind.
White, J. A., Box 1230, Williamstown, Mass.
White, J. R., 33 Hidden Lane, Westbury, L. I., N. Y.
Whiting, Webster K., 604 Woodlawn Ave., Falls Church, Va.
Whitney, D. J., 41 Gabb Rd., Bloomfield, Conn.
Whittaker, H. F., Jr., 32 Essex St., Lynn, Mass.
Widenor, M. D., 19 Greely Sq., Glen Head, L. I., N. Y.
Widmann, F. W., 15 Roberts Ave., Haddonfield, N. J.
Wier, J. M., 904 S. Wabash, Urbana, Ill.
Wiesner, L., 118—65 Metropolitan Ave., Kew Gardens, L. I., N. Y.
Wild, J. H., 1224 Ransom St., Sandusky, Ohio
Wild, J. J., 7 Overhill Lane, Roslyn, L. I., N. Y.
Wilkinson, D. A., 777—14 St., N. W., Washington, D. C.
Willard, J. M., 5205 Eight Rd., So., Arlington 4, Va.
Willard, R. S., Jr., 40 Kaiholu Pl., Lanikai, Oahu, T. H.
Willett, E., 447 Tenth Ave., New York 1, N. Y.
Williams, C. E., 1517 Piedmont Ave., Austin, Tex.
Williams, C. H., 2031—34 St., S.E., Washington 20, D. C.
Williams, C. S., 5010 Kimball, Kansas City 4, Kan.
Williams, J. T., 145 Callan Ave., Evanston, Ill.
Williams, R. J., 111, Greene Rd., Berwyn, Pa.
Williams, R. W., 4800 Oak Grove Dr., Pasadena, Calif.
Williams, S. B., 4847 N. 41 St., Milwaukee 9, Wis.
Williamson, C. R., 85 E. Columbus Ave., White Plains, N. Y.
Williamson, J. C., 205 Brow St., Liverpool, N. Y.
Williamson, R. J., 136 Esperanza Ave., Sierra Madre, Calif.
Willis, F. H., Bell Tel. Labs., New York 14, N. Y.
Willner, P., 499 Fort Washington Ave., New York, N. Y.
Willson, F. E., 463 West St., New York 14, N. Y.
Wilmot, R. D., Kirk's Ferry, Ottawa, P.Q., Canada
Wilson, C. A., 1612 Cody Dr., Silver Spring, Md.
Wilson, H. R., 236 S. Lynnwood Ave., Glenside, Pa.
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Wingate, Sidney A., 8 Gray St., Cambridge, Mass.
Winget, William F., 2342 Commonwealth Ave., St. Paul 8, Minn.
Winkle, La Verne, 1270 Prospect Pl., Cincinnati 31, Ohio
Winkler, M. R., 5225 N. 20 St., Phoenix, Ariz.
Winter, Harold F., Mountain Ave., Murray Hill, N. J.
Withrow, W. E., 4803 Ravenswood Rd., Riverdale, Md.
Wittenberg, R. C., 136 Pine St., New Hyde Pk., L. I., N. Y.
Wittig, William L., 2519 S. Walter Reed Dr., Arlington 6, Va.

(Continued on page 125A)

Membership

(Continued from page 124A)

Wohl, C. J., 44-21 Macnish St., Elmhurst 73, L. I., N. Y.
Wolf, A. A., 7131 Cedar Park Ave., Philadelphia 38, Pa.
Wolf, E. W., 19 Robinson Rd., Lexington, Mass.
Wolf, F. E., Jr., 161-03—65 Ave., Flushing, L. I., N. Y.
Wolfand, D., 905 Manor Rd., Falls Church, Va.
Wolfe, C. M., 820 E. Bennett, Glendora, Calif.
Wolken, I. I., 6307 N. Francisco Ave., Chicago 45, Ill.
Wong, G. W., 66 University Rd., Brookline 46, Mass.
Wood, Frederick B., 1152 Hedding St., San Jose, Calif.
Wood, R. M., Jr., 13 Gibson Rd., Hampton, Va.
Wood, R. F., Bear Rd., N. Syracuse, N. Y.
Woodruff, A. E., 312 N. Parkside, Chicago, Ill.
Woodruff, R. L., 86 Concord St., San Francisco 12, Calif.
Woods, O. R., 301 Henrietta St., Kalamazoo, Mich.
Woods, W. A., 255 Waterview St., Playa Del Rey, Calif.
Woodward, W. E., 5002 S.E., 122 St., Portland 66, Ore.
Woodyard, J. R., University of California, Berkeley, Calif.
Worthen, G. B., 2630 Kingsbridge Terr., Bronx, N. Y.
Wrathall, G. R., Rt. 2, Box 260, Aptos, Calif.
Wright, C. R., 207 St. Lo., Mishawaka, Ind.
Wright, D., 1420 Opechee Way, Glendale 8, Calif.
Wright, D. S., 86 Brook Rd., Red Bank, N. J.
Wright, J. W., 5504 Grove St., Chevy Chase, Md.
Wright, R. B., 4607 Connecticut Ave., N.W., Washington, D. C.
Wright, T. A., Highland Pk., Village, Dallas, Tex.
Wurman, G., 318 E. Walnut St., Long Beach, L. I., N. Y.
Wycoff, K. H., Rt. 1, Overton, Nebr.
Wyeth, F. H., 4901 Stenton Ave., Philadelphia 44, Pa.
Yang, T., Remington Rand Inc., Philadelphia, Pa.
Yaplee, B. S., 4219—Fourth St., S.E., Washington 20, D. C.
Yarbrough, A. D., 3513 Stanford St., Dallas 5, Tex.
Yeager, J. R., Queens Village, L. I., N. Y.
York, E., Jr., 1094 S. King St., San Gabriel, Calif.
Young, A. F., Rt. 2, Box 474-A, Severna Pk., Md.
Young, J. W., Box 407, Fort Amador, Balboa, Canal Zone
Young, M. F., Box 1142, Honolulu, T. H.
Young, W. E., Hq. Fifth Air Force, Box 544, APO 710, c/o P.M., San Francisco, Calif.
Youngblood, W. A., M.I.T., Cambridge 39, Mass.
Younger, U. E., 3202 E. Foothill Blvd., Pasadena 8, Calif.
Youngstrom, N. C., Seven E. Cedar Lane, Maplewood, N. J.
Youtz, P., 211 Massachusetts Ave., Cambridge 39, Mass.
Yu, Y. P., 451 Highland Ave., Passaic, N. J.
Zablocki, H. S., 30 Grove Terr., Irvington, N. J.
Zancanata, H. W., R.F.D. 2, Aberdeen, Md.
Zavac, F. R., Rockaway Dr., R.D. 2, Boonton, N. J.
Zebrowitz, S., 2220 Pine St., Philadelphia 3, Pa.
Zeiger, L. B., 4515—12 Ave., Brooklyn 19, N. Y.
Zenon, Z., 27 Lincoln Dr., Box 363, N. Syracuse, N. Y.
Zern, R. T., 2972 Second St., Norristown, Pa.

(Continued on page 126A)

Membership

(Continued from page 125A)

Zieman, C. M., 2293 Hazelton Ave., Dayton 3, Ohio
Zilis, G. S., 6742 S. Artesian Ave., Chicago 29, Ill.
Zimmerman, C. W., 3326 E. Haskell Pl., Tulsa, Okla.
Zimmerman, J., 1950 N. River Rd., Melrose Pk., Ill.
Zimmerman, R. L., 17422 Osborne, Northridge, Calif.
Zuckerberg, H., 67-30 Dartmouth St., Forest Hills, L. I., N. Y.
Zuckerbraun, J., 2260 Olinville Ave., New York, N. Y.

Admission to Member

Adams, J. E., Box 138, North Bay, N. Y.
Adams, R., Suite 1, 38 Home, Winnipeg 10, Man., Canada
Baron, J., 3 Rue J. R. Bloch, Sartrouville (S&O) France
Barre, W. E., Sylvania Electric Products, Inc., 12 Second Ave., Warren Pa.
Bateman, F. F., 2500 Morton St., Charlotte, N. C.
Bly, J. R., 4831 Angeles Crest Hgwy., La Canada, Calif.
Bower, C. M., Air Associates, Inc., 511 Joyce St., Orange, N. J.
Callahan, A. L., U.S. Embassy, APO 928, c/o P.M., San Francisco, Calif.
Chaudhuri, D., 500 Riverside Dr., New York 27, N. Y.
Chu, Y., 9402 Allen Rd., Allen Park 10, Mich.
Cleveland, A. R., 11413 Lanewood Cir., Dallas, Tex.
Cook, M. E., 534 Baeder Rd., Jenkintown, Pa.
Cory, W. E., 254 Shrine, San Antonio, Tex.
Durage, A. H., 70 Wootton Rd., Essex Falls, N. J.
Farrell, J. W., 1536 San Saba Dr., Dallas 18, Tex.
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de Haro, F. O., Yerbal 2663, Buenos Aires, Argentina
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(Continued on page 130A)

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(Continued from page 128A)

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(Continued on page 132A)

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(Continued from page 130A)

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Fig. 1 — *Plate current plotted against plate voltage for one triode section of a 12AU7. Plate load is 5 k, peak plate-supply voltage is 500 v. Grid voltage is changed 5 v between curves, from —35 v. to zero. Vertical sensitivity is 5 ma/div, horizontal sensitivity 50 v/div. Calibrated controls permit accurate current and voltage readings directly from the screen.*

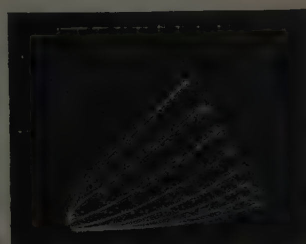

Fig. 2 — *Same triode section of 12AU7 with only 20-v peak plate supply and sensitivities increased to 0.2 ma/div vertical and 2 v/div horizontal. Grid voltage is changed 2 v between curves, from —14 v to zero. This is essentially a 25-times magnification of the lower left portion of Fig. 1, showing the operating characteristics at low plate-supply voltage.*

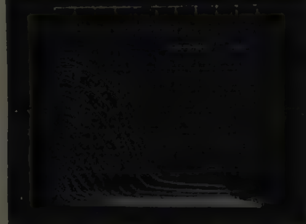

Fig. 3 — *Screen current plotted against plate voltage with positive grid bias on a 6AQ5. Plate load is 300 ohms, peak plate voltage is 100 v, screen-grid voltage is 100 v, with grid voltage changing 2 v/step from +16 v to below zero. Vertical scale is 10 ma/div, horizontal scale 10 v/div.*

Fig. 4 —*Typical Germanium Diode curve. Inherent flexibility of the Type 570 permits accurate evaluation of diode characteristics and detailed examination of any part of the curve. Calibrated scales above are 0.2 v/div horizontal, 0.5 ma/div vertical, with zero points at center of screen.*

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Industrial Engineering Notes*

FCC Actions

The Federal Communications Commission has issued a request to set manufacturers that they supply information relating to the current status of UHF television receivers. The request was contained in a notice of proposed rule making which looks toward increasing the maximum power limits for UHF stations to five million watts. In other actions the commission issued rules authorizing the construction of low-power TV stations, without regard to city size, and amended the chain broadcast rules to prevent territorial exclusivity clauses in TV station contracts with networks.

Mobilization

The Office of Naval Material has made a report available in which it showed that 509 manufacturers of electronic end equipment and major components had sales of nearly $6.6 billion in 1954. These firms employed over 458,000 persons in connection with their electronic end equipment production and have a maximum annual production potential, on a one-shift basis, of over $9.6 billion.

RETMA

Dorman D. Israel of Emerson Radio & Phonograph Corp. has been named Chairman of the Joint Technical Advisory Committee for a one-year period starting July 1. The appointment was made by W. R. G. Baker, Director of the RETMA Engineering Department. Ernst Weber of the Polytechnic Institute of Brooklyn has been appointed Vice-Chairman for the same period by the Board of Directors of the Institute of Radio Engineers. The other members of JTAC for the coming year are R. N. Harmon, Westinghouse Broadcasting Co., Inc.; I. J. Kaar, Gen. Elec. Co.; A. V. Loughren, Hazeltine Electronics Corp.; Ralph Bown, Bell Telephone Labs.; John V. L. Hogan, Hogan Laboratories, Inc., and Philip F. Siling, Radio Corp. of America. L. G. Cumming, Technical Secretary of the IRE, will continue to serve as Secretary. . . . **W. R. G. Baker, Director of the RETMA Engineering Department, has been elected to the grade of Fellow in the Standards Engineering Society.** The award will be made at an awards luncheon Sept. 30 during the group's national convention at the Hotel Statler in Hartford, Conn.

Standardization

The Radio Technical Commission for Aeronautics has released a report setting forth minimum performance standards for airborne radio communication receiving

(Continued on page 136A)

* The data on which these Notes are based were selected by permission from *Industry Reports*, issues of June 13, and 26, July 4, and 11, published by the Radio-Electronics-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

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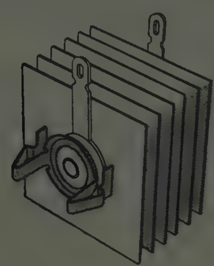

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Industrial Engineering Notes

(Continued from page 134A)

equipment operating in the 118-132 meg. band. Compliance with the standards is recommended by RTCA as a means of assuring "that the equipment will satisfactorily perform its intended function under all conditions normally encountered in routine aeronautical operations." In addition to setting forth minimum performance standards, the report also outlines standard test conditions and methods of test to be used in measuring performance characteristics. "Minimum Performance Standards—Airborne Radio Communication Receiving Equipment Operating Within the Radio-Frequency Range of 118–132 Megacycles" is available at 30 cents per copy from the RTCA Secretariat, Room 2036, Building T-5, Washington 25, D. C.

Technical

A summary of the ideas and experiences of researchers in the field of automatic programming for digital computers, compiled by the Office of Naval Research, is contained in a report made available to industry by the Office of Technical Services, Commerce Department. The volume consists of papers relative to differentiators, compilers, generators, interpretative routines and universal automatic codes. The report, PB 111607, "Symposium on Automatic Programming for Digital Computers," may be obtained from OTS, Commerce Department, Washington 25, D. C., for $4. Also, OTS has made available "A survey of High-Speed Printers for Digital-Computer Output," prepared in August 1952 by the Office of Naval Research. The code number of this report is PB 111615, and the price is 50 cents. . . . **An improved sphere generator for precision contouring of round quartz crystals on a mass production basis, developed by Bausch and Lomb under an Army Signal Corps contract, is described in a report just released to industry by the Office of Technical Services, Commerce Department.** The report notes that the contouring machine grinds spherical bevels on one or more flat round crystals at a time, with a range of curves from 29 mm to 350 mm. Curves even shorter can be generated with slight modifications in the machine and its auxiliary equipment, the OTS report states. The report, "Contouring Equipment for Round Crystals," is the final report on the subject, and can be ordered by number—PB 111609—from OTS, Commerce Department, Washington 25, D. C., for $1.50 each, which includes drawings and design data. . . . **The Radio Technical Commission for Aeronautics recently released two reports relative to airborne radio communication equipment.** The first report sets forth minimum performance standards for airborne radio communication transmitting equipment operating within the radio-frequency range of 118–132 mc. A similar report dealing with receiving equipment in this band was released previously (RETMA Industry Report, Vol. 11, No. 24). The RTC report, "Minimum Per-

(Continued on page 138A)

Eimac Amplifier Klystrons and Circuit Components

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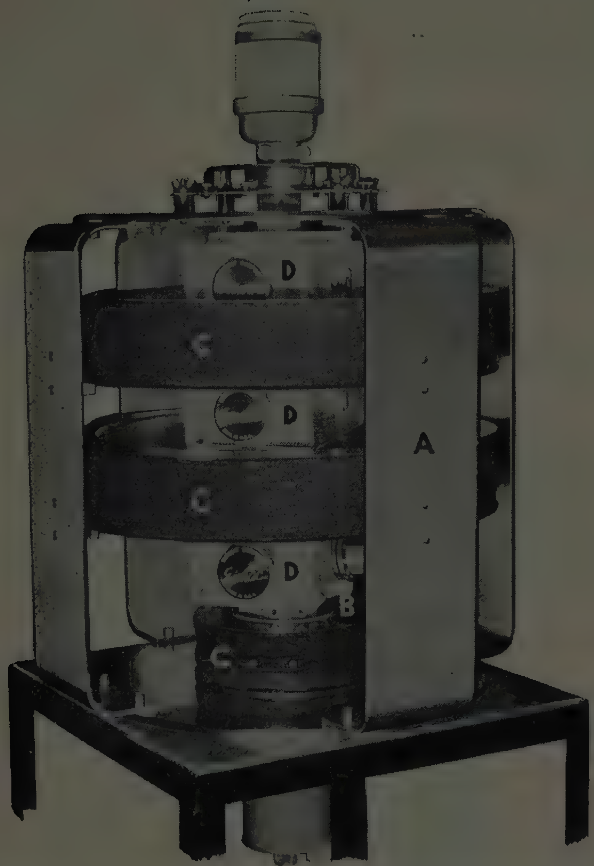

Design and construction of a high power UHF/microwave transmitter for beyond-the-horizon communication and other microwave applications is simple and straight-forward with an Eimac amplifier klystron and circuit components. In fact, it's easier to build than a low frequency Class C amplifier. Eimac high power amplifier klystrons, plus Eimac circuit components consisting of A) Magnetic framework B) RF output load coupler C) Magnetic beam-control coils and D) Convenient tuning wide range RF resonant cavities comprise the essential elements of a final amplifier package. By adding conventional power supplies, control circuits, driver and cabinets to the Eimac klystron-circuit component package, high power at UHF is easily obtained. Eimac developed klystron and circuit components provide equipment manufacturers with the easy economical approach to high power microwave transmitters. In many cases, existing low power equipment can be used as a driver for the higher powered amplifier.

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C
Magnetic beam-control coils

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Resonant cavities

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	FREQUENCY RANGE—MC	CW POWER
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3K20,000LK	720-890	5000w
3K50,000LA	470-580	10,000w

	FREQUENCY RANGE—MC	CW POWER
3K50,000LF	580-720	10,000w
3K50,000LK	720-890	10,000w
3K50,000LQ	850-1000	10,000w
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Industrial Engineering Notes

(Continued from page 136A)

formance Standards, Airborne Radio Communication Transmitting Equipment Operating Within the Radio-Frequency Range 118–132 Megacycles," is available at 30 cents a copy from the RTCA Secretariat, Room 2036, Building T-5, Washington 25, D. C. Compliance with the standards, the report stated, is recommended as a means of assuring that the equipment "will satisfactorily perform its intended function under all conditions normally encountered in routine aeronautical operations." The second RTCA report issued, "Re-Evaluation of VOR Airway Lateral Separation Criteria" discusses the problems of VOR stations and receivers and describes methods of procedure for solving the problems. The procedures described are those relating to service and laboratory tests. The text also contains a statistical analysis of VHF omni-range navigation receiver errors and an analysis of VHF omni-bearing system errors. The report is available at 75 cents a copy from the RTCA Secretariat. . . . **Two manuals designed to assist manufacturers of electronics equipment for the Armed Forces were just issued by the Office of Technical Services, Department of Commerce.** One manual is designed to help the manufacturer meet the requirements of radio-interference specifications and the other is concerned with the application of electron tubes. "Radio Interference Suppression Techniques" contains information on approved suppression components and systems and their application. In addition, it explains the procedures recommended for obtaining approvals, requesting tests by the Signal Corps, and in getting assistance from the Signal Corps in solving special problems. Some of the individual sources of radio-interference dealt with are rotating machinery, ignition systems, switches and contactors, electronic devices, fluorescent lamps and instruments. The manual is available from the OTS, Commerce Department, Washington 25, D. C., for $6.75 and should be ordered by number—PB 111611. "Techniques for Application of Electron Tubes in Military Equipment" presents tube information from the point of view of the electronic design engineer, and is organized under three sections: numerical data and special design considerations for specific tube types; tube properties according to ratings, characteristics essential in circuit operation and properties detrimental to circuit operation; and tube properties in circuit design including a check list for the use of the circuit designer to insure coverage of all important design factors. The concepts of specification control, operation within ratings, and tolerance characteristic are emphasized throughout the report, available from the OTS, Commerce Department, for $2.50. The report should be ordered by number—PB 111644. . . . **The Office of Technical Services, Commerce Department, has announced studies in the field of electronics in its "U. S. Govern-**

(Continued on page 142A)

Industrial Engineering Notes

(Continued from page 138A)

ment Research Reports." The following reports can be purchased from the Photoduplication Section, Library of Congress, Washington 25, D. C., for the reported price: "Analysis of a Cyclotron Type Tube," PB 116027; microfilm, $3; photocopy, $7.75."Circular Waveguide Coupling for Laboratory Use," PB 116623; microfilm, $1.50; photocopy, $1.50. "Coaxial Line Step Discontinuity Admittances," PB 116547; microfilm, $2; photocopy, $2.75. "Accelerated Life Program," PB 116407 (quarterly progress report No. 1) and PB 116409 (quarterly progress report No. 3); microfilm, $2.25; photocopy, $4. "Radio Interference Suppressors. Final Report Under Contract No. AF 33 (038)-9353," PB 116519; microfilm, $4.75; photocopy, $14. "Flush-mounted Cardioid-Pattern Antennas," PB 116070; microfilm, $2.50; photocopy, $5.25. "Integral Equation Governing Electromagnetic Waves," PB 116389; microfilm, $2; photocopy, $2.75. "Presentation on Automatic Production of Electronic Equipment," PB 116381; microfilm, $3; photocopy, $7.75. "Measurement of Pulse Time Jitter," PB 116561; microfilm, $2.25; photocopy, $4. "On the Cylindrical Antenna," PB 116304; microfilm, $2.25; photocopy, $4. "Shielded Two-Wire Hybrid Junction," PB 116292; microfilm, $2.25; photocopy, $5.25. "Semi-Flush-Mounted Cardioid-Pattern Antenna for the 225–400 mc Band," PB 116071; microfilm, $2; photocopy, $2.75. "State of the Art of Electronic Subminiaturization," PB 115555; microfilm, $6.75; photocopy, $22.75. "Transmission-Line Tubes," PB 116537; microfilm, $2.25; enlargement print, $6.50.

Television

"Approximately 50,000 color television sets will be produced this year and it will be six or seven years before color set production catches up with black-and-white," Board Chairman Max F. Balcom estimated during a speech at the Canadian RETMA Convention in Niagara Falls, June 2–3. Mr. Balcom discussed "The Status of Color Television in the U. S." and told the Canadians that "by the end of the decade it is likely that 7.4 million sets will be sold, at a factory price total of $1.5 billion." He pointed out that a number of factors are involved in the growth of color TV and "it is not simply a case of designing a color set and rushing it out to the market and handing it to some eager customer who will pay anything he has to pay to get it" and ". . . at this stage of the game, color television is filled with unpredictables and uncontrollables. . . . As far as some specific estimates about color production and sales are concerned" he said, "our latest thinking is that some 50,000 sets will be produced this year in the U. S. and that something less than 35,000 will probably be sold at the retail level. You will recall that last year some 25,000 sets were produced but indications are that not more than 10,000 of them were sold to the

(Continued on page 154A)

Industrial Engineering Notes

(Continued from page 150A)

CHEMISTS, ENGINEERS, PHYSICISTS, PATENT AGENTS

Florida firm with Manhattan office has several positions available for research and development on radioactive batteries. Vacuum evaporation experience is helpful but not essential. Radiation Research Corp., 140 Cedar St., New York 6, N.Y.

ENGINEERS

We want experience in magnetic amplifiers, analog computers, servomechanisms and electronic systems. Juniors to $120.00 per week. Intermediates to $140.00 per week. Box 825.

PATENT RESEARCH ENGINEER

Engineering Degree, Electrical or Electronics major preferred for opening involving technical research and report preparation based on aeronautic applications. Opportunity for an engineering graduate interested in entering the patent field. Liberal salary commensurate with experience and educational qualifications. Pension plan. Nationally known organization located in New York City. Box 826.

MISSILES SYSTEMS ENGINEERS

For work on radar indicator circuits, digital computers, systems analysis, missile planning. Submit inquiries to Engineering Personnel Division, Sperry Gyroscope Co., Great Neck, L.I., N.Y.

ASSISTANT OR ASSOCIATE PROFESSOR

Ph.D. preferred to teach undergraduate and evening graduate courses. Specialist in Communication or Computers preferred. Salary on 9 month basis competitive, with opportunity for local summer employment for right man. High altitude, dry climate. Write Chairman, Elect. Engr. Department, University of New Mexico, Albuquerque.

TECHNICAL SALES ENGINEER

A medium-sized southern New England manufacturer of technical papers used by the electronics industry has a challenging position for an electrical engineer under 35. Some selling or administrative experience is desirable. He must have a strong persuasive personality and be capable of working with both technical and non-technical people. Salary open. Include resume of personal history, education, and experience with introductory letter to Box 827.

ELECTRONIC ENGINEERS

Must have 2 to 6 years experience in the design of I.F., V.H.F., and/or U.H.F. broad band systems. Progressive L.I firm; many benefits Send resume or phone for appointment: Mr. G. W. Fellendorf, PIoneer 2-5300, Instruments for Industry, Inc., 150 Glen Cove Road, Mineola, N.Y.

ASSISTANT PROFESSOR

Assistant professor of Electrical Engineering, University of North Dakota, Grand Forks, North Dakota. Duties: teaching electronic and communications courses to undergraduate students. An advanced degree plus some teaching or industrial experience preferred. Submit resume of education and experience to Head, Electrical Engineering Dept.

ELECTRONICS ENGINEERS

Degrees in electrical or electronic engineering with at least 2 years related experience, to conduct and supervise junior engineers and technicians on qualification testing and evaluation of electronic components. Analyze test programs and instrumentation, design test methods, evaluate completed data and prepare final reports. Age; 25 to 40. U. S. Testing Company, Inc., 1415 Park Avenue, Hoboken, N.J.

violin music

The demand exceeds the supply. It's that simple! With 3 engineering jobs available for every 2 engineers, some 5,000 companies are bidding for you with offers, inducements and background music. But don't be mistaken! Most of today's "opportunities" are *jobs*, not futures.

We, too, want engineers. But we're offering no violin music—only the opportunity for intelligent and careful evaluation—you of us and we of you—with the possibility of your joining one of the finest team operations in the whole new world of flight systems development.

Most of the people on that team are young, and moving ahead fast. They weren't lured here. They found out—and figured out—for themselves. We hope you'll do that too.

Write to J. M. Hollyday, Dept. P-9, The Glenn L. Martin Company, Baltimore 3, Maryland.

MARTIN
BALTIMORE · MARYLAND

Industrial Engineering Notes

(Continued from page 142A)

public, since they were really developmental and prototype models. Making any predictions about next year is really not advisable until we can get a more definite measure of this year, but today's commonly accepted figure is that some 150,000 sets will be made in 1956. Several years from now, and on into 1960 or so, it is probable that about one-third of all sets produced will be color sets, and that color will be nearing two-thirds of all sets in nine or ten years. In other words, it will be six or seven years before color set production catches up with black-and-white." . . . **A committee of engineers appointed by the Senate Interstate and Foreign Commerce Committee to study the present allocation of the spectrum and other engineering matters related to TV (RETMA Industry Report, Vol. 11, No. 25) held its first meeting and named Professor Edward Bowles of the Massachusetts Institute of Technology as permanent Chairman.** Donald Fink, of Philco Corp., representing RETMA, was named Secretary. In addition to Edward Bowles and Donald Fink, other members of the engineering committee are: Allen B. DuMont and Robert Wakeman of Allen B. DuMont Laboratories, Inc.; radio engineering consultants C. M. Jansky and S. Bailey; William Duttera, National Broadcasting Co.; Frank Marx, American Broadcasting Co.; William Lodge, Columbia Broadcasting System; Ralph Harmon, Westinghouse Broadcasting Co., Inc.; Haraden Pratt, Institute of Radio Engineers; Curtis Plummer, Federal Communications Commission, and engineering consultant T.A.M. Craven.

IRE People

Consolidated Engineering Corporation has appointed **R. B. Hurley** (S'44-A'47), a transistor specialist, to its Research Division as a senior research engineer.

R. B. Hurley

Prior to his appointment by Consolidated, Mr. Hurley spent three years as senior electronics engineer with Convair, Pomona. Earlier, he was an engineer with Aerojet-General Corporation, Azusa, California; U. S. Electrical Motors, Los Angeles, and General Electric Company, Schenectady, New York. He is also a part-time lecturer in engineering at the University of California in Los Angeles.

He is author of two articles on transistor amplifiers which appeared recently and "Predictable Design of Transistor Ampli-

(Continued on page 156A)

IRE People

(Continued from page 154A)

fiers," will appear this fall in Tele-Tech & Electronic Industries.

Holder of the M.S. degree in electrical engineering from the University of Southern California, Mr. Hurley is a member of Phi Beta Kappa, Eta Kappa Nu, Sigma Xi, Tau Beta Pi, American Institute of Electrical Engineers, and the Scientific Research Society of America.

❖

The appointment of **T. R. Kennedy Jr.** (SM'48), to the public relations staff of Allen B. Du Mont Laboratories, Inc., has been announced.

For twenty-five years Mr. Kennedy was a member of the radio-television news and editorial staff of The New York Times. He began newspaper work in 1925 at the Pittsburgh Post & Sun, being radio editor and radio advertising manager. A year later he joined the Philadelphia Evening Ledger as technical radio editor.

Mr. Kennedy studied electrical engineering at Carnegie Institute of Technology, Pittsburgh, and joined the U. S. Navy at the beginning of World War I. He attended the Navy Radio School at Harvard University, received a commission, and was transferred to the Boston Navy Yard, where he helped design, construct, and install the radio compass.

Mr. Kennedy is a member of the Audio Engineering Society, a fellow of the Radio Club of America, and a member of the Radio Pioneers.

❖

Lester M. Field (M'48–F'52) has been appointed Head of the Electron Tube Laboratory, Hughes Aircraft Company Culver City, California. He comes to Hughes with a background in tube research and development acquired at the Bell Telephone Laboratories and as head of tube research projects at both Stanford University and the California Institute of Technology. He will continue to be associated with the faculty at the California Institute of Technology to help maintain the research program in progress there.

Dr. Field's honors include one of the Eta Kappa Nu Outstanding Young Electrical Engineer Awards for 1949, and election to Fellow of the Institute of Radio Engineers. He is also a member of Tau Beta Pi, Sigma Xi, the American Association of University Professors, and the American Physical Society.

❖

J. W. Thatcher (SM'46), formerly an engineering executive at Western Electric Company, has been appointed Customer Service Manager of Pasadena's Electro-Data Corporation.

Mr. Thatcher will head a department of computer engineers and technicians,

(Continued on page 161A)

IRE People

(Continued from page 156A)

responsible for nationwide installation and maintenance of "DATATRON" high-speed electronic data processing machines.

During 26 years with Western Electric, Mr. Thatcher has held a variety of posts, most recent of which was the west coast managership of field engineering service for electronic equipment purchased by the Air Force. Previously he was west coast representative of the Nike contract manager. He has served also as manager of the Wesco Field Modification Shop and general manager of Sound Services, Inc., a Western Electric subsidiary.

He is a graduate of the California Institute of Technology.

❖

The election of **Richard Hodgson** (M'47–SM'51) as a Vice-President of Fairchild Camera and Instrument Corporation has been announced. At the same time, Mr. Hodgson was promoted from the position of Trend Planning Director to that of General Manager of the Company's Reconnaissance Systems Division.

Mr. Hodgson was one of the organizers of Chromatic Television Laboratories, Inc., which pioneered in the development of manufacture of color television picture tubes based on the invention of Dr. E. O. Lawrence. He was president of the firm when he joined Fairchild, and continues to serve on the board of directors.

His association with the military dates back to 1942 when he was a research staff member of M.I.T.'s Radiation Laboratory working on microwave radar developments. His work included liaison with Air Force Laboratories at Wright Field on airborne radar developments and operational applications. In 1944 he was given leave and assigned to the office of the Secretary of War as an expert consultant on radar. During this period he also served as civilian radar advisor to General Hoyt B. Vanderburg, then Commanding General of the 9th Air Force.

He also served as a consultant to the Department of the Air Force in 1952 and 1953, advising on the organization of the research and development program as part of a three-man advisory group headed by Lt. General James H. Doolittle. This work led to the establishment of the Air Research and Development Command.

He also has been Director of Television Development for Paramount Pictures Corporation; Assistant Treasurer of Allen B. DuMont Laboratories, Inc.; head of the Engineering Management Division of Brookhaven National Laboratory, A.E.C.; a senior change board engineer for Lockheed Aircraft Corp. and a manufacturing and process economic analyst for Standard Oil Company of California.

He is a member of the Society of Motion Picture and Television Engineers and has served on many industry committees of the Radio-Eletronics-Television Manufacturers Association.

(Continued on page 162A)

IRE People

(Continued from page 161A)

The new Technical Products Division of Allen B. DuMont Laboratories, Inc., will be headed by **P. S. Christaldi** (S'35–A'40–SM'44–F'52) and will manufacture and sell the products formerly handled by the company's Instrument Division and Communication Products Division. It is initiating a program of expansion and diversification with especial emphasis on elements and systems for automation of industrial processes.

P. S. CHRISTALDI

Products manufactured by the new division include television station broadcasting and studio equipment, industrial television systems, cathode-ray and other electronic instruments, automotive test equipment, industrial electronics, components and systems, guided missiles test equipment, military electronic and electromechanical devices, and the newly developed complement of mobile radio equipment.

Dr. Christaldi has been associated with Allen B. DuMont Laboratories since 1938. His first duties were in the field of cathode-ray tube and cathode-ray oscillograph development, being expanded to include television receiving and transmitting operations when he was appointed Chief Engineer of the company in 1941. In 1947 he became Engineering Manager of the Instrument Division. He was made Assistant Manager of the division in 1952 and Manager in 1953. He is a member of Sigma Xi, the American Radio Relay League, the American Rocket Society, a Fellow of the Radio Club of America, and a fellow of the Institute of Radio Engineers. Dr. Christaldi is a member of the IRE Standards Committee, Chairman of the IRE Committee on Measurement and Instrumentation, a member of the IRE Subcommittee on Oscillography, chairman of the RETMA Committee on Oscillography, and IRE Delegate to the American Standards Association, Section C-39.

❖

Franklin S. Cooper (A'46) research scientist and physicist, has been elected President and Director of Research of Haskins Laboratories Inc. of New York to succeed Caryl P. Haskins who has resigned to accept the presidency of the Carnegie Institution of Washington. Mrs. Anne Hinchey Gallagher was elected to serve as Treasurer and Assistant Secretary.

Franklin Cooper was born in Robinson, Illinois in 1908 and graduated from the University of Illinois. He received the

(Continued on page 165A)

IRE People

(Continued from page 162A)

Ph.D. degree in 1936 from the Massachusetts Institute of Technology. He was with the Research Laboratory of the General Electric Company until he joined Haskins Laboratories in 1939 where he has been Associate Director of Research.

Dr. Cooper was Senior Liaison Officer of the Office of Scientific Research and Development during the war, in charge of the exchange of scientific information with the Allies. He has served as consultant to the Secretary of Defense and to the Atomic Energy Commission of the United Nations. He is a member of various scientific and professional societies and of advisory committees of the Massachusetts Institute of Technology and New York University, and is Adjunct Professor of acoustic phonetics at Columbia University.

❖

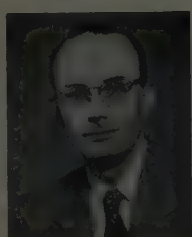
The appointment of **W. Evan-Jones** (A'54) to the Canadian Westinghouse Company Ltd., Electronics Division in Hamilton, Ontario has been announced. Mr. Evan-Jones will be responsible for the Sales Engineering promotion of the company's communication products across Canada.

Prior to joining the Canadian Westinghouse Company, Mr. Evan-Jones was with the Canadian Marconi Company as Assistant Sales Supervisor, Commercial Products Division Montreal; The Communications Group, Hydro Electric Power Commission of Ontario; and the Northern Electric Co. Ltd.

He has been actively engaged in communication engineering for the past fifteen years and is an Associate Member of the American Institute of Electrical Engineers.

❖

Airborne Instruments Laboratory, Inc., Mineola, New York, announces the promotion of **J. G. Stephenson** (A'45–M'47–SM'50) to the position of assistant Supervising Engineer of the Applied Electronics Section of the Research and Engineering Division.



J. G. Stephenson

In the ten years of his employment with Airborne Instruments Laboratory, he has been associated with major developments in the fields of UHF receivers, stable oscillators, and data-processing equipment.

Prior to his employment with Airborne Instruments Laboratory in 1945, Mr. Stephenson was a Research Associate at the Radio Research Laboratory of Harvard University, where he worked on

(Continued on page 166A)

IRE People

(Continued from page 165A)

radar countermeasures. He also served with the American British Laboratory in England and was a Technical Observer in the Mediterranean Area.

Mr. Stephenson graduated from Yale University with the B.E. degree in 1939 and from Stanford University in 1941 with the Engineer degree. He is a member of the AIEE, Sigma Xi, and Tau Beta Pi.

❖

Edward H. Weitzen, president of the Gruen Watch Company, has announced the appointment of **Gerald C. Schutz** (M'46–SM'50) as Director of Electronics.

G. C. SCHUTZ

Mr. Schutz was formerly associated with the Gibbs Manufacturing and Research Corporation as Director of Electronics. During his five years with the Gibbs Organization, he was responsible for the initiation of a number of development programs in the airborne radar navigation and countermeasures field.

Previously, Mr. Schutz was in charge of Airborne Radar Techniques development for the U. S. Air Force at Wright Field, Dayton, Ohio. With the Air Force from 1942 to 1950, he was concerned with the development of many radar test instruments including radar missile guidance and bombing systems.

Mr. Schutz is a graduate of the University of Illinois with the Bachelor of Science degree in electrical engineering. He is a member of Tau Beta Pi.

❖

P. J. Schenk (A'43–SM'52), former executive officer to the Technical Assistant to the Secretary of the Air Force, has been appointed manager of marketing for the General Electric Light Military Electronic Equipment Department.

P. J. SCHENK

His new duties will include direction of the department's product planning, sales, contract negotiation, and product and field service activities.

Mr. Schenk for the past year has been manager of market research and product planning for the G. E. Heavy Military Electronic Equipment Department, Syracuse, N. Y.

Mr. Schenk, a Lieutenant Colonel in the USAF Reserve, joined G. E. in early 1954 following his Air Force assignments

(Continued on page 167A)

IRE People

(Continued from page 166A)

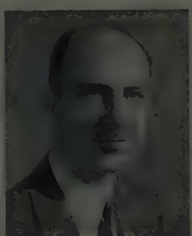
as executive officer to Lt. Gen. James H. Doolittle, technical assistant to the Secretary of the Air Force; assistant military director of the Air Force Scientific Advisory Board; and executive officer to the Chief Scientist, USAF.

From mid-1951 to December, 1952, he served successively as deputy for research and as vice commander of the Air Force Cambridge Research Center, Cambridge, Mass. Previously, he served for two years on the Research and Development Planning Team of the office of the Deputy Chief of Staff, Development, USAF. From 1946 to 1949 he was chief, Aircraft Control and Warning Branch, Headquarters, USAF, and from 1943 to 1946 he was technical inspector and deputy signal officer of the 26th Fighter Command in Panama. He entered active duty with the Signal Corps in 1941, and for two years instructed and organized in the radio and radar school at Ft. Dix, N. J. He has a mobilization assignment to the Scientific Advisory Board, Office of the Chief of Staff, Headquarters, USAF.

A native of Vienna, Austria, Mr. Schenk came to this country with his parents at the age of 12 and settled in New York City. He won a four-year regional competitive scholarship to Lafayette College, Easton, Pa., and was graduated in 1941 with honors.

❖

Administration of technical product sales for Allen B. DuMont Laboratories. Inc., will be handled by **G. Robert Mezger**, (A'37–VA'39–SM'54), who has been named General Sales Manager of the Technical Products Division.



G. R. Mezger

Mr. Mezger will supervise the Television Transmitter Sales Department the Mobile Communications Sales Department, and the Technical Sales Department. He will be responsible for over-all sales planning, administration, and sales promotion for DuMont electronic and cathode-ray instruments, television transmitting and studio apparatus, mobile radio equipment, industrial cathode-ray tubes, and multiplier phototubes, electronic control apparatus, and industrial electronic devices.

Immediately prior to his new appointment, Mr. Mezger was Assistant Manager of the Instrument Division.

He first joined the DuMont organization in 1936, following his graduation from Rensselaer Polytechnic Institute. From 1936 to 1939 he was active in the engineering and development of cathode-ray instruments and was technical sales manager from 1939 until 1941, when he was assigned to active duty by the United States Navy. With the Navy he participated in instrument development work at the David

(Continued on page 169A)

I.R.E. People

(Continued from page 167A)

Taylor Model Basin in Washington, D.C. and in the design of naval radar equipment. Mr. Mezger held the rank of Commander. He rejoined DuMont after World War II where he served successively as technical sales manager, engineering manager, and assistant division manager—all for the Instrument Division.

He is a member of the American Institute of Electrical Engineers and the Institute of Radio Engineers.

❖

M. J. Kelly (M'25–F'38), president of Bell Telephone Laboratories, New York, has been named to the Board of Trustees of Stevens Institute of Technology, Hoboken, N. J.

M. J. KELLY

Dr. Kelly has served as chariman of the U. S. Air Force's Scientific Advisory Board, and as advisor to the Secretary of the Air Force, the Atomic Energy Commission and other defense agencies.

An alumnus of the Missouri School of Mines, he received the master's degree from the University of Kentucky and the Ph.D. from the University of Chicago.

He is a Fellow of the American Physical Society, the Acoustical Society of America, and the American Institute of Electrical Engineers; and a member of the National Academy of Sciences, the American Philosophical Society, and Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

❖

The appointment of **Peter Janis** (SM'47) as Chief Engineer of Amperex Electronic Corporation, Hicksville, L.I., New York has been announced.

PETER JANIS

Mr. Janis came to Amperex from Sylvania Electric Products, Inc., where he was active in the development of klystrons, hydrogen thyratrons, magnetrons and photo-sensitive devices. Later, he was put in charge of the microwave and special purpose tube development activities of the Product Development Laboratories, and more recently was involved in the design, development, and production of traveling-wave tubes.

Prior to his work at Sylvania, he was engaged at R.C.A., Victor Division in development of special purpose tubes. He has also taught theory and design of monochrome and color television, AM and FM receivers.

Born in New York, Mr. Janis received

(Continued on page 170A)

IRE People

(Continued from page 169A)

the E.E. degree from Cooper Union School of Engineering, and attended Columbia University and New York University. He is a recipient of the War Production Merit Award, has several patents and disclosures on electronic devices, and has written articles and papers on wide band oscillators, amplifiers and microwave devices.

Section Meetings

ALBUQUERQUE-LOS ALAMOS

"A Transistorized Digital Computer — TRADIC," by J. H. Felker, Bell Telephone Labs.; July 14, 1955.

ATLANTA

"Radio Location by Phase Comparison," by E. J. Crossland, Seismograph Corp.; June 24, 1955.

(Continued on page 172A)

Section Meetings

(Continued from page 170A)

BALTIMORE

"Ferrites in Duplexers," by T. N. Anderson, Airtron, Inc.; June 8, 1955.

BEAUMONT-PORT ARTHUR

"The IRE," by D. J. Tucker, Director, Region 6; June 21, 1955.

CEDAR RAPIDS

"Basic Considerations and More Interesting Problems in the Development of the Iowa Electronic Test Processing Equipment," by John A. Brady; "The Future of the Peaceful Atom," by Aristids Ratermanis; and "Matrix Analysis of Communication Networks and Power Systems," by D. R. Wilson; May 18, 1955.

"Elements of Music and an Electronic Music Synthesizer," by Dr. H. F. Olson, RCA Labs.; June 15, 1955.

CINCINNATI

"Jest about Jets," by N. R. Thomas, General Electric Company; June 21, 1955.

DENVER

Business meeting, and inspection trip to TV transmitter sites of KLZ-TV, KBTV and KOA-TV; June 7, 1955.

DETROIT

Student papers contest: "Putting the Rainbow on the Air," by R. P. Farbolin, Wayne University; "Semi-Conductors and Point Contact Transistors," by Elliot Rappaport, Wayne University; "Magnetic Amplifiers Using Square Loop Core Materials," by J. W. Rood, Michigan State University; May 20, 1955.

EL PASO

"Electronics in Law Enforcement," by Lt. Art Clatfelter, El Paso Police; June 30, 1955.

FORT WAYNE

"Galactic Radio Waves," by Dr. J. P. Hagen, U. S. Naval Research Lab.; June 2, 1955.

HAWAII

"1955 IRE Convention Report," by J. R. Sanders, Matson Navigation Co.; July 13, 1955.

HUNTSVILLE

"Radio Astronomy," by H. W. Wells, Carnegie Institute and Director, Region 3; June 23, 1955.

INDIANAPOLIS

"Radar Systems, All Types and Their Characteristics," by C. J. Marshall, Wright Patterson AFB; June 9, 1955.

INYOKERN

"Modern Radar Components and Applications," by George Dexter, Hoffman Labs.; July 11, 1955.

OKLAHOMA CITY

Tour of the International Crystal Company Plant; April 20, 1955.

Student papers: "Electrical Pipeline Pumping," by Bob McAlpine; "The Economical Replacement of Machinery," by Jess Jackson; "Power Plant Location," by Bob Marshall; and "Something for Nothing," by Ted Hart; May 4, 1955.

PITTSBURGH

"Numerical Control," by H. P. Grossimon, Massachusetts Institute of Technology; June 13, 1955.

SACRAMENTO

Business and social meeting, with talk by J. H. Cowan, former member of the F.B.I.; June 17, 1955.

(Continued on page 174A)

Section Meetings

(Continued from page 172A)

San Antonio

"Instrumentation for Uranium Prospecting," by Dr. J. C. Cook, Southwestern Research Institute; April 27, 1955.

Upgrading party; May 27, 1955.

"Telemetering Highlights of 1955," by John Ohman, Southwest Research Institute; June 3, 1955.

San Francisco

"The Engineer in Society," by J. W. McRae, Sandia Corporation; June 17, 1955.

Seattle

Student paper contest: "Amplifiers," by Phil Beaudoin; "An Investigation of the Two-Charge Theory of Electrets," by R. B. Kieburtz; and "Dielectric Potentiometers," by A. T. Simmons, all of University of Washington; April 12, 1955.

"Automatic Electronics," by J. D. Ryder, President, IRE; April 21, 1955.

"Instrumentation at the Hanford Operation," by R. G. Clark, and "A Gamma Scintillation Spectrometer," by R. E. Connally, both of Hanford Works; May 19, 1955.

Election of officers; June 3; 1955.

Toledo

"Police Radar, Speed Control," by Manford Rosencrantz, Toledo Police Dept.; April 14, 1955.

Technical Tour of Television Facilities at WSPD-TV, conducted by W. M. Stringfellow; May 12, 1955.

Election of officers; June 9, 1955.

Vancouver

"Automatic Electronic Production," by J. D. Ryder, President, IRE; April 20, 1955.

"Atomic Energy—The New Frontier," by Dr. G. M. Shrum, U.B.C.; May 19, 1955.

SUBSECTIONS

Buenaventura

"The Present and Future of Airborne Fire Control," by Dr. R. M. Ashby, North American Aviation, Inc.; June 9, 1955.

Tucson

"Applied Communication Theory," by Dr. Eberhardt Rechtin, California Institute of Technology; April 22, 1955.

"Automation and Mechanization," by G. O. Haglund, General Mills; May 26, 1955.

Wichita

"Boolean Algebra and Its Circuit Application," by P. I. Miller, Boeing Airplane Co.; June 2, 1955.

Election of officers; June 17, 1955.

Kay Lab. Industrial TV Demonstration in Mobile Trailer, by Mr. O'Neal and Bob Bacon, Both of Kay Lab.; June 23, 1955.

Professional Group Meetings

Aeronautical and Navigational Electronics

Dallas-Ft. Worth Chapter—May 24

"Modern Aircraft Transmitter Design" by Ryan B. Seals, Collins Radio Company.

Professional Group Meetings

New York Chapter—May 19

"Distance Measuring Equipment" by William J. Flynn, Hazeltine Electronics Corporation.

"Automatic Test Equipment for DME" by Sherman Rinkle, Polytechnic Research and Development Company, Inc.

Election of Officers: Chairman—Gordon P. McCouch, Vice-Chairman—Lester M. Glantz, Secretary—William P. McNally.

Antennas and Propagation

Albuquerque-Los Alamos Chapter—May 4

"Theoretical Analysis of a Dipole Antenna" by S. H. Dike, Sandia Corporation.

Chicago Chapter—April 15

"Ionospheric Propagation" by Howard J. Goldman, Armour Research Foundation.

Chicago Chapter—May 20

"A Color Film Entitled '*The Antenna is the Payoff*'" by The Channel Master Corporation of Ellenville, New York.

Audio

Boston Chapter—June 2

"Man, a Somewhat Neglected Component of Hi-Fi Systems" by Walter A. Rosenblith, MIT.

Chicago Chapter—January 19

"Calibration of Test Records by Interference Patterns" by Ben Bauer, Shure Brothers, Inc.

Chicago Chapter—April 15

"The Circulatron Amplifier" by John Overly and Lloyd Loring, Electro-Voice.

Cincinnati Chapter—October 19

"Audio Systems Home Demonstration" by C. G. Haehnle, A. B. Bereskin, F. M. Livezey, J. P. Goode, R. A. Jenkins, J. P. Quitter, W. W. Gulden.

Cincinnati Chapter—November 16

"What and How You Hear" by Warren E. Johnson, Cincinnati Speech and Hearing Center.

Cincinnati Chapter—January 18

"A Twin Lever Ceramic Cartridge" by B. B. Bauer, Shure Brothers, Inc.

Cincinnati Chapter—April 26

"How Many Dollars Can You Hear?" by Harvey B. Glatstein, Customcrafters Audio, Inc.

Cincinnati Chapter—May 17

"Binaural Audio Demonstration" by C. G. Haehnle, E. D. Kay, and R. A. Jenkins, Jr.,

Syracuse Chapter—April 13

"Performance of the 'Distributed Port' Loudspeaker Enclosure" by A. F. Petrie, Gen. Elec. Co.

(*Continued on page 176A*)

Professional Group Meetings

(Continued from page 175A)

Broadcast and Television Receivers

Chicago Chapter—April 15

"A New High Efficiency Parallax Mask Color TV Tube" by Mark Amdursky, Rauland Corp.

Chicago Chapter—May 20

"Theory and Design of Vertical and Horizontal Deflection Circuits" by Kurt Schlesinger, Motorola, Inc.

Broadcast Transmission Systems

Houston Chapter—May 30

"Color TV Demonstration of '*The Petrified Forest*'."

Circuit Theory

Albuquerque-Los Alamos Chapter—May 25

"Physics of Transistors" by Walter E. Brown, Sandia Corporation.

Election of Officers: Chairman—Norman J. Elliott, Vice Chairman—George Webber, Secretary—John McLay.

Chicago Chapter—April 15

"Effects of Impedance in Cathode and Plate of Vacuum Tube Amplifiers" by Bernard S. Parmet, Motorola, Inc.

Los Angeles Chapter—May 13

"Color Television Circuitry" by Dr. E. L. Michaels, Packard-Bell Corp.

"Color Television Studio Equipment, Staging Practices, and Techniques" by W. H. Copeland, CBS.

Election of Officers: Chairman—J. E. Jacobs, Secretary-Treasurer—H. Low, Program Committee—J. Heilfron, R. M. W. Johnson, Dick Towle.

Urbana Chapter—February 10

"Formulation of Kirchhoff Voltage and Current Postulates" by M. B. Reed and E. W. Schwarz, Univ. of Illinois.

Urbana Chapter—February 24

"The Differential Equation Postulates for Circuit Theory" by M. B. Reed and E. W. Schwarz, Univ. of Illinois.

Urbana Chapter—March 10

"Mathematical Aspects of Switching" by F. E. Hohn, Univ. of Illinois.

Urbana Chapter—March 24

"The Differential" by M. E. Munroe, Univ. of Illinois.

Urbana Chapter—April 14

"The Stieltje's Integral" by F. E. Hohn, Univ. of Illinois.

Urbana Chapter—April 28

"The Laplace-Stieltje's Transform" by S. Seshu, Univ. of Illinois.

Urbana Chapter—May 12

"Time Domain Analysis" by C. L. Coates, Univ. of Illinois.

(Continued on page 178A)

Professional Group Meetings

(Continued from page 176A)

Urbana Chapter—May 26

"Purpose of Symbolic Logic" by H. E. Vaughan, Univ. of Illinois.

Electronic Computers

Boston Chapter—May 26

"Pulse Circuits Near Absolute" by Dudley A. Buck, M.I.T.

Dallas-Ft. Worth Chapter—March 22

"Introduction to Control Systems Application" by John F. Nolan, M.I.T.

"Development of a Magnetic Memory" by Harlan E. Anderson, M.I.T.

New York Chapter—April 26

"Fundamental Theory and Operation of a Typical but Elementary Computer (Digital Differential Analyzer)" by J. A. Githens, Bell Labs.

"Design Features of the ETT-100 Computer, (Stevens D.D.A.)" by S. M. Shockell, Stevens Institute.

New York Chapter—May 24

"The Type 650 Magnetic Drum Data Processing Machine" by R. W. Avery, IBM Corp.

Electronic Devices

Los Angeles Chapter—May 2

How Should a Transistor be Characterized?"——Panel Discussion.

Philadelphia Chapter—May 17

"Transistors Today" by Harry L. Owens, Signal Corps Engineering Labs.

Election of Officers: Chairman—Bernard T. Svihel, Vice Chairman—Werner Hasenberg, Secretary—Gordon R. Spencer.

Washington, D. C. Chapter—May 23

"P-N-I-P Transistors" by J. M. Early, Bell Labs.

Election of Officers: Chairman—Henry D. Arnett, Vice Chairman—Robert W. Grantham, Secretary—Harold J. Peake.

Engineering Management

Los Angeles Chapter—May 18

"A Technique for Interviewing" by Dick S. Barlow, Hughes Aircraft Co.

Philadelphia Chapter—May 26

"Research and Development Organization and Planning for a Dynamic Field" by J. W. Forrester, Lincoln Computer Labs., also Director, Digital Computer Lab., MIT.

Syracuse Chapter—May 12

"Planning Meeting for Engineering Management Course" by Harlan Perrins, Cornell Univ.

Professional Group Meetings

Industrial Electronics

Chicago Chapter—April 15

"Automation in the Electronics Industry" by V. C. Lafferty, Armour Research Found.

Cleveland Chapter—April 21

"Radio Noise Measurements and Control Problems" by Leonard Thomas, Bureau of Ships, Navy Dept.

Election of Officers: Chairman—Ruben Kozarian.

Information Theory

Albuquerque-Los Alamos Chapter—May 11

Business Meeting

Election of Officers: Chairman—Walter E. Brown, Vice-Chairman—Leo V. Skinner, Secretary—Charles H. Bidwell.

Los Angeles Chapter—May 26

"Linear Programming Applied to Bid Analysis" by Leon Gainen, Hughes Aircraft Co.

Instrumentation

Houston Chapter—May 24

"Nuclear Batteries" by Edward A. DeCrosta, Tracerlab, Inc.

Medical Electronics

Los Angeles Chapter—January 26

"Medical and Electronic Cooperation" by J. Phillip Sampson, L. A. County Medical Assoc.

"Instrumentation Problems" by Marcel Verzeano, UCLA.

Los Angeles Chapter—March 16

"Heart Microphone" by John Hilliard, Altec Lansing.

"X-Ray Microscope" by Curtis G. Smith, UCLA.

Los Angeles Chapter—April 19

"Hearing Correction" by Helen Kennedy, City College.

"Electric Potentials in Tissue" by Robert Tschirgi, UCLA, "Electrosurgery" by Ralph Gunter, UCLA, "Physiological Monitoring" by John Dillon, UCLA.

Los Angeles Chapter—May 10

"Acoustical Resonance in Microorganisms" by Robert Woods, College of Med. Evang.

"Localization of Nerve Blocks" by Robert Pearson, College of Med. Evang.

San Francisco Chapter—May 26

"Electronic Instrumentation in Otology and Audiology" by Francis Sooy, Univ. of Cal., and R. E. Allison, Allison Labs.

Microwave Theory and Techniques

Boston Chapter—May 12

"Broadband Microwave Crystals" by Eugene J. Feldman, Sylvania Elec. Products Inc.

(Continued on page 180A)

Professional Group Meetings

(Continued from page 179A)

Chicago Chapter—April 15

"Latest Developments in Microwave Test Equipment Featuring the Reflectometer" by John E. Stiles and Frank Waterfall, Alfred Crossley Associates, Inc.

Nuclear Science

Albuquerque-Los Alamos Chapter—May 3

Nomination of Officers for coming year: Chairman—Alan S. Rawcliffe, Vice-Chairman—Leo Grant, Secretary—Richard Hiebert.

Albuquerque-Los Alamos Chapter—May 18

"The Racetrack Synchrotron" by G. P. Kenfield, Sandia Corp.

Boston Chapter—April 14

"Dynamic Analysis and Control of Nuclear Power Plants" by J. N. Grace, Westinghouse Atomic Power Div.

Election of Officers: Chairman—John C. Simons, Vice-Chairman and Program Chairman—A. B. Van Rennes, Secretary—William M. Trenholme.

Chicago Chapter—Arpil 15

"Methods of Particle Acceleration" by John J. Livingood, Argonne National Lab.

Chicago Chapter—May 20

"Pioneering in Nuclear Research" by Hal Fredrich Fruth, Hal Fruth Associates.

Connecticut Valley Chapter—May 31

"The Atomic Energy Industry" by F. H. Warren, Gen. Dynamics Corp.

Election of Officers: Chairman—Nelson A. Merritt, Vice Chairman #1—Francis A. Fanelli, Vice Chairman #2—John E. Bassett, Secretary-Financial Officers—Henry A. Sardelli.

Oak Ridge Chapter—May 18

"Organization and Programming for Magnetic Drum Calculators" by J. P. Kelly, K-25 Plant, Oak Ridge.

Production Techniques

Washington, D. C. Chapter—April 14

"The Autofab Assembly System" by Donald Melton, General Mills.

Reliability and Quality Control

Chicago Chapter—May 20

"Reliability of Transformers" by Martin Wolff, Chicago Transformer Corp.

Telemetry and Remote Control Communications Systems

Chicago Chapter—April 15

"A New Data Transmission System for Pipeline Applications" by George E. Foster, Metrotyne Corp.

Chicago Chapter—May 20

"The Role of Magnetic Tape in Data Recording, Processing and Analysis" by Gomer L. Davies, The Davies Lab.

(Continued on page 181A)

Professional Group Meetings

(Continued from page 180A)

Los Angeles Chapter—May 17

"Linear Voltage-Controlled Frequency Modulation of the Hartley Oscillator" by W. F. Link, Bendix Aviation Corp. (Presented by W. G. Coe, Bendix Aviation Corporation).

"High Measurement Accuracy via FM/FM Telementry" by D. W. Blancher, Bendix Aviation Corp.

VEHICULAR COMMUNICATIONS

Chicago Chapter—April 15

"Transistorized Auto Receiver" by Thomas O. Stanley, RCA Labs.

News–New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 16A)

TV Camera Dolly

A new, lightweight, remarkably maneuverable television camera pedestal is introduced by **Houston Fearless Div., Color Corp. of America,** 11831 W. Olympic Blvd., Los Angeles 64, Calif. Designated the PD-7, it is particularly adapted to the small studio, or is suitable as an auxiliary camera mount for larger studios.

Outstanding feature of the Houston Fearless PD-7 Pedestal is its unusually light weight: 140 lbs. It rolls smoothly on rubber-tired ball bearing wheels and will pass through a 34 inch or, if necessary, a 30 inch door.

Its maneuverability is made possible by two types of steering: For rolling dolly shots, "parallel steering" is used wherein the three

(Continued on page 183A)

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For the expansion of a small group of competent physicists and engineers who are concerned with the development of new devices and with the solution of advanced instrumentation and measurement problems. This group is responsible for devising methods for the solution of special problems and for the experimental verification of these methods. The final engineering and packaging is normally carried out by other groups in the organization. The varied nature* of this work requires both recent graduates and experienced people capable of accepting primary responsibility for the solution of problems of varying degrees of complexity.

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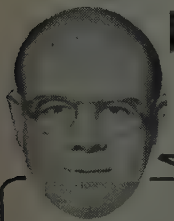

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(Continued from page 181A)

wheels are locked in parallel and turn together in any direction. For sharp turning or rotating "tricycle steering" is employed wherein all steering is done with the rear wheel while the front wheels are locked in parallel. Changing from one type of stecring to the other is done simply by lifting the steering wheel.

RF Signal Generator

EICO ELECTRONIC INSTRUMENT CO., INC., 84 Withers St., Brooklyn 11, N. Y., has just released details on its new Model 324 Signal Generator. It can be used for IF-rf alignment; signal tracing, and trouble shooting of AM, FM and TV receivers (all on fundamentals); as a marker generator for alignment of new high-frequency as well as older low frequency TV IF's; for 400 cps sine wave audio testing, and laboratory and experimental work.

Its frequency range is 150 kc to 145 mc on fundamentals in 6 bands; and 111 mc to 435 mc on calibrated harmonics. Accuracy of dial calibration is ± 1–5 per cent with 6–1 vernier tuning know and good spread at most important frequencies.

Colpitts rf oscillator, directly plate-modulated by a cathode follower for improved modulation. Turret-mounted, slug tuned coils for individual calibration of each band. Variable depth of internal modulation from 0 to 50 per cent by the 400 cps Colpitts oscillator. Variable gain external modulation amplifier requires audio input of only 0.8 volts for 30 per cent modulation Fine and Coarse (3 step) attenuators—50 ohms output impedance. The unit is supplied complete or in kit form. Request catalog Z-1.

(Continued on page 184A)

News–New Products

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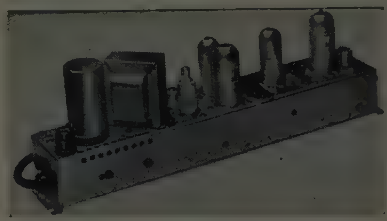
(Continued from page 183A)

Master TV Amplifier with AGC

High power single channel TV amplifiers with automatic gain control are now available from **Blonder-Tongue Laboratories, Inc.**, 526–536 North Ave., Westfield, N. J. These model MCS units are used to amplify, mix and equalize individual VHF channels at the antenna site, or in TV distribution lines.



The new "Masterline" amplifiers deliver a maximum constant output of 0.7 volts RMS on each channel. Gain ranges from 32 to 36 db and response is flat within ±1 db. Side channel rejection is high. Any number of units may be mixed without loss and channels may be cascaded for increased gain. Model MCS can be added to any existing amplifier system.

Each MCS has a built-in power supply and an ac receptcble to receive the power cord of the next unit. Power rating is 28 Watts at 0.28 amperes from a standard 117 volt source. Tube complement includes (2) 6CB6, (1) 6BK7A and (1) 6AM8. All input and output cables are handled by 75 ohm coax fittings. In outdoor installations, amplifiers may be enclosed and shielded in B-T's Model MRH Radiation-Proof Housing. List price for each MCS unit is $109.50.

Current Probe

A new device for the measurement of rf current and interference in conductors has been announced by the **Stoddart Aircraft Radio Co., Inc.**, 6644 Santa Monica Blvd., Hollywood 38, Calif.

The probe does not require direct connection to the conductor under measurement. It is a special-

(Continued on page 186A)

E-cores, L-cores, and U-cores

Deflection yoke cores Quarter-section

Toroid, cylindrical, and ring cores

FERRITE COMPONENTS of HIGH EFFICIENCY for COLOR TV CIRCUITS

A greatly broadened line of Allen-Bradley Quality ferrite parts is now available to electronic and television set manufacturers. Some standard pieces are shown above.

Three performance standards—WO-1, WO-2, and WO-3 have been established for the electrical and magnetic characteristics of Allen-Bradley ferrite component parts:

WO-1 and WO-3 are somewhat more efficient but still interchangeable with other makes of ferrite parts.

WO-2 parts have much lower losses and higher permeability with greater flux density at maximum operating temperatures. Their higher magnetic efficiency permits reduction in size of these ferrites and the use of less copper. A lower over-all cost is often the result. In some color television circuits, the use of Allen-Bradley WO-2 ferrites has eliminated two tubes and related parts.

Allen-Bradley has grown rapidly as a dependable producer of Quality ferrite parts. It will pay you to investigate the performance of Allen-Bradley ferrites in your electronic circuits.

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OTHER QUALITY COMPONENTS FOR RADIO, TV & ELECTRONIC APPLICATIONS

NEW — Allen-Bradley Ferrite Beads for increasing the inductance of lead wires.

ALLEN-BRADLEY FERRITE BEADS

Ferri-Cap Feed-thru Filters are capacitors in combination with ferrite material to provide "T" filter performance.

ALLEN-BRADLEY FERRI-CAP FILTERS

ALLEN-BRADLEY

RADIO, ELECTRONIC AND TELEVISION COMPONENTS

News-New Products

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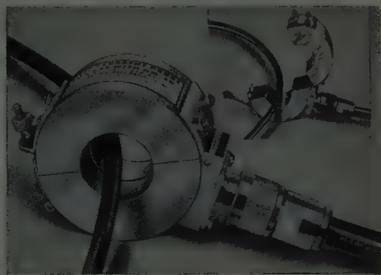
(Continued from page 184A)

ly-designed radio-frequency current transformer of the inserted-primary type and has a nominal output impedance of 50 ohms. The probe consists of two semi-circular insulated windings on a hypersil core. These are hinged, and by opening the probe the conductors may be placed in its center. A locking arrangement holds the probe closed.



The probe is useable from 20 cps to 25 mc and is especially designed for use with the Stoddart NM-10A and NM-20B Radio Interference-Field Intensity Measuring equipments.

Interference measurements may be made on single and multi-conductor cables, on ground and bonding straps, and on the external surfaces of shielding conduits and coaxial cables. Radio frequency currents, modulated or unmodulated, can also be measured by suitable operation of the NM-10A or NM-20B.

Precision Variable Delay Line

This delay line, designed by **Advance Electronics Co., Inc., 451** Highland Ave., Passaic, N. J., consists of 60 sections of LC m-derived networks and one 60 position rotary switch. The LC m-derived networks are especially designed for fast rise time, and negligible overshoot. The rotary switch is used to change the amount of time delay between the input and output by connecting the output terminal to any one of the 60 sections of LC networks. The input impedance of the delay line is equal to the characteristic impedance. The output terminal should be connected to a high-impedance load (approximately 20 times or more the characteristic impedance of the line), such as the

(Continued on page 188A)

News-New Products

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(Continued from page 186A)

grid circuit of the amplifier. The end of the delay line has been terminated by a resistor equal to the characteristic impedance internally.

Both the m-derived networks and the rotary switches can be moved from the cabinet and incorporated into any equipment where various time delay is needed. The accuracy of time delay can be as high as ± 0.5 per cent of the time delay at any point.

There are five different types available. Type 605a has a maximum time delay of 0.6 μs in step of 0.01 μs, 75 ohms impedance, 32 megacycles bandwidth, and 0.01 μs rise time. Type 605b has a maximum time delay of 1.5 μs in step of 0.35 μs, 95 ohms impedance, 12.4 mc bandwidth, and 0.025 μs rise time. There are three more types available with total time delay up to 12 μs. All of them have 60 steps and impedance ranging from 75 ohms to 300 ohms.

Miniature Toroid Coils

A line of toroid coils for use with transistors in both subminiature and printed circuits is announced by the **Forrest Mfg. Co.,** 5962 Smiley Dr., Culver City, Calif. These coils are designed for insertion through transistor terminal borads, the windings are terminated in lugs, or solid wire leads. Leads can be tinned for dip soldering, or brought to any pin arrangment. Major applications for the toroids include uses in

(Continued on page 190A)

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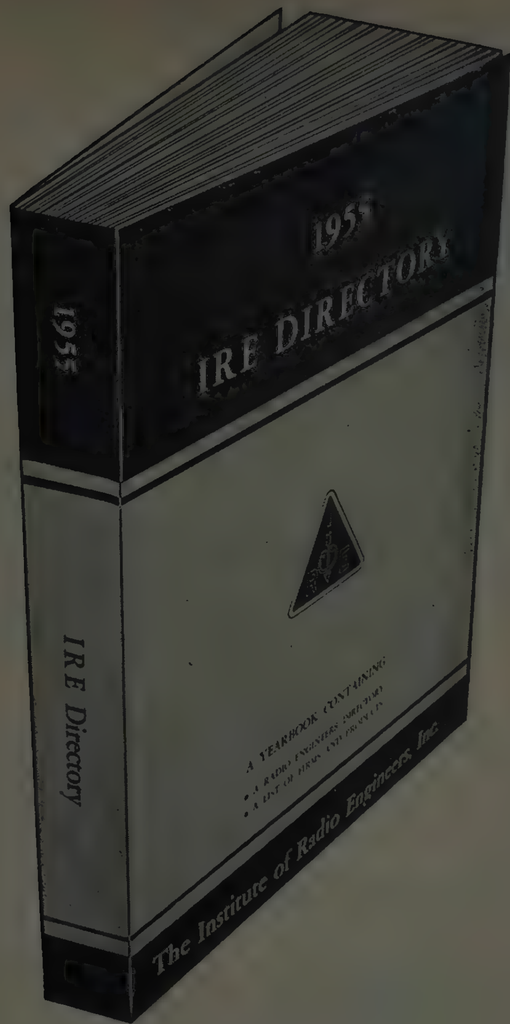

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News–New Products

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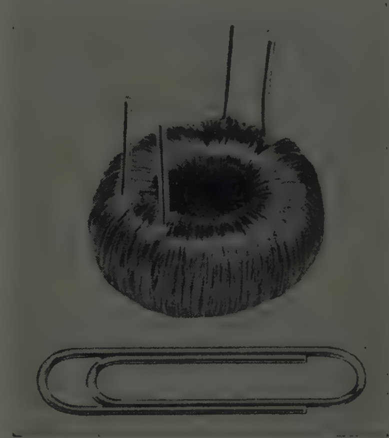
(Continued from page 188A)

super-selective IF, rf and audio transformers, pulse transformers, high Q filter reactors, magnetic amplifiers, low-level modulators, tuneable reactors, and voltage regulators.



Sizes range from $\frac{5}{8}$ OD$\times\frac{1}{4}$ to $\frac{7}{8}$ OD$\times\frac{3}{8}$ inches. Inductance values up to 2 henries. Types for use with reactors provide as high as 150 Q in the $\frac{5}{8}$ inch size illustrated, with rising Q values as coil sizes increase. Types for use in magnetic amplifiers, modulators and voltage regulators absorb up to 130 V, 400 cps in units with an OD of 0.94 inch. Manufacturer states that windings with low capacity and rf resistance can be provided. Units are finished with hard-setting epoxy resin, or are sealed in cans to meet Mil specifications. Maker will wind coils to special sizes and inductance values. Details on the line will be sent on request.

Color TV Tubescope

The Color TV Tubescope, manufactured by **Edmund Scientific Corp.**, 101 E. Gloucester Pike, Barrington 3, N. J., is a microscope designed to assist and simplify alignment of the dot pattern on the picture tube during manufacture and on regular field service calls. Specifically, it checks the accuracy and concentricity of the pattern where unaided vision is unsatisfactory.

The instrument consists of two parts, a main chrome plated optical body housing a cemented achromatic objective lens with the eyepiece, and the outer black

(Continued on page 192A)

News–New Products

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(Continued from page 190A)

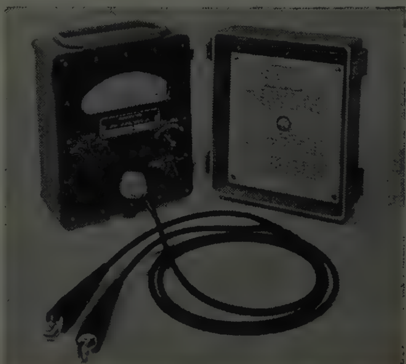
anodized felt lined aluminum barrel for focusing. A working distance of $4\frac{1}{4}$ inches is ample to view the periphery of the tube, and

makes it unnecessary to remove the safety glass shield thereby protecting technicians from the high potential on the tube.

Five different eyepieces are available and can be easily interchanged to give powers from 10× to 33×. The instrument is $7\frac{1}{4}$ inches long with outside diameter of $1\frac{1}{2}$ inches.

Ohmmeter

A new low-resistance ohmmeter featuring low test current is announced by **Industrial Instruments, Inc.,** Cedar Grove, N. J., manufacturers of electrical test equipment. The new meter, Model LRO, is especially adaptable to measurement of relay contact resistance, fuse resistance, bonding and ground wire resistance and any other application requiring low resistance measurements with minimum test current.



The Model LRO has an accuracy of ±1 per cent of full scale reading. Ranges include full-scale

(Continued on page 194A)

new

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smaller · more efficient
minimum radio noise
MEETS MIL-M-8609 SPECS

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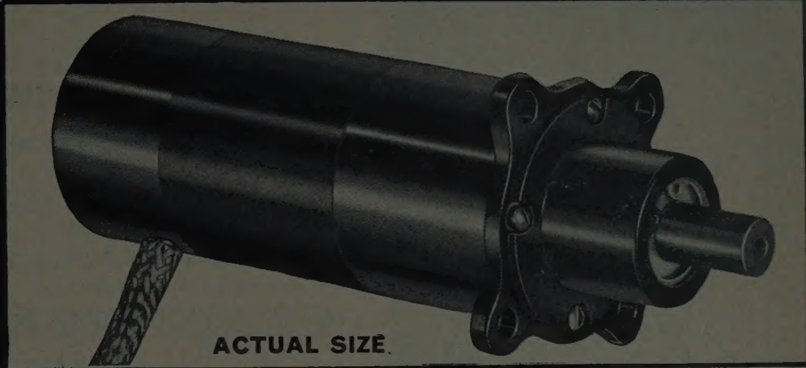

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- Available with gear train, governor, brake or any combination thereof. For gear train ratios, see chart.
- Applications: radio, radar, actuators, drive mechanisms, antenna tilt-motors, tuning devices, blowers, cameras and many others.

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PERMANENT MAGNET MOTOR GEAR TRAIN DATA

TORQUE AT OUTPUT SHAFT OZ. IN.	GEAR RATIO OF GEAR TRAIN
25	15:1 to 33,000:1
100	15:1 to 33,000:1
300	15:1 to 33,000:1
400	15:1 to 5,500:1
600	15:1 to 5,500:1

Motor can be designed for speeds from 6000 RPM to 20,000 RPM.
Length of motor will vary according to power.
Length of gear train will vary according to gear ratio required—

1000:1 to 33,000:1	6 stages
300:1 to 5,900:1	5 stages
100:1 to 1,000:1	4 stages
40:1 to 183:1	3 stages
15:1 to 32:1	2 stages

Other products include Actuators, AC Drive Motors, DC Motors, Fast Response Resolvers, Servo Torque Units, Servo Motors, Synchros, Reference Generators, Tachometer Generators and Motor Driven Blower and Fan Assemblies.

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(Continued from page 192A)

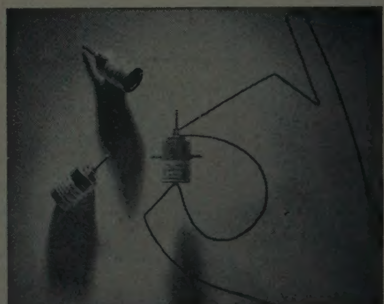
readings of 0.1 ohm, 1 ohm and 10 ohms. The scale is divided into 1100 divisions, that is, one scale division on the 0.1 range therefore being 0.001 ohm.

With the Model LRO Ohmmeter current passing through test piece never exceeds 110 milliamperes on any range. Protection for the meter is incorporated in the event resistance being measured is outside range of instrument or open. Connection to resistance to be measured is made by special test lead supplied with the instrument which provides four terminal measurement for increased accuracy in the low resistance ranges.

Instrument is entirely self-contained. Power is supplied by standard flashlight battery of $1\frac{1}{2}$ volts.

Variable Inductors

Available both shielded and unshielded, then ew Style A Type 1 Variable Inductors designed by **Levinthal Electronic Products, Inc.,** 2979 Fair Oakes Ave., Redwood City, Calif, are supplied in ten standard values from 56 μh to 1.8 μh and up to a maximum of 25 μg in special units. They feature an inductance variation range of 2:1; Q's of approximately 200; operating temperature range from -50°C to $+100$°C; and temperature coefficients of inductance less than 50 ppm/°C.



Imbedment of the entire powdered carbonyl-iron cup core and coil assembly in epoxy resin gives these variable inductors high resistance against the effects of large-amplitude vibration or shock, as well as providing good protection against moisture and chemical attack. Type 1 core formula gives unusually constant

(Continued on page 196A)

COMMUNICATIONS EQUIPMENT CO.

MICROWAVE COMPONENTS

10 CM.—RG48/U Waveguide

POWER SPLITTER for use with type 726 or any 10 CM Shepherd Klystron. Energy is fed from Klystron antenna through dual pick-up system to 2 type "N" connectors $12.50

LHTR. LIGHTHOUSE ASSEMBLY. Parts of RT39 APG 5 & APG 15. Receiver and Trans. Cavities w/assoc. Tr. Cavity and Type N CPLG. To Recvr. Uses 2C40, 2C43, 1B27, Tunable APX 2400-2700 MCS. Silver Plated $15.00

BEACON LIGHTHOUSE cavity p/o UPN-2 Beacon 10 cm. Mfg. Bernard Rice, each $27.50

MAGNETRON TO WAVEGUIDE Coupler with 721-A Duplexer Cavity, gold plated $31.50

721A TR BOX complete with tube and tuning plungers $12.50

McNALLY KLYSTRON CAVITIES for 707B or 2K28, 2700-2900 MC $4.00

HOLMDELL-TO-TYPE "N" Male Adapters, W. E. $D167284 ....$2.75

BEACON ANTENNA AS31/APN-7 in Lucite Ball, Type "N" feed $22.50

ANTENNA, AT49A/APR: Broadband Conical, 300-3300 MC Type "N" Feed $12.50

"E" PLANE BENDS, 90 deg. less flanges $7.50

K-Band, X-Band Eqpt. Available........Send for List

X BAND—1" x ½" WAVEGUIDE

PARABALOID DISH, 18" diam. Spun Aluminum, 8" Focus. For AN/APS-6 ....$4.95

3 CM. DIPOLE and Feed Assembly. (May be used with above dish.) 8 inches long $5.00

FLEXIBLE SECTION 9 in. long, Cover-to-Cover $5.50

ROTARY JOINT (APS-6) Sperry PT $658275, 180 deg. rotation, choke to choke. Has "Built-in" Di-Coupler, 20 DB., with "N" Takeoff $22.50

3CM. DIPOLE FEED, 15" L. for APS-15 ......$14.50

MITRED ELBOW, Cast aluminum, 1¾" x ⅝" W.G. W.E. Flanges. "E" Plane $3.50

3 CM ANTENNA ASSEMBLY: Uses 17" paraboloid dish, operating from 24 vdc motor. Beam pattern: 5 deg. in both Azimuth and elevation. Sector Scan: over 160 deg. at 35 scans per minute. Elevation Scan. over 2 deg. Tilt. Over 24 deg..$35.00

Cross-Guide Directional Coupler, UG-40 output flange. Main Guide is 6" Long, with 90 Deg. "E" Plane bend at one end, and is fitted with Std. UG 39/UG 40 flanges. Coupling figure: 20 db Nominal $22.50

RG52/U Waveguide in 5' lengths, fitted with UG 39 flanges to UG40. Silver plated ......**per length $5.00**

Rotating-Joints supplied either with or without deck mountings. With UG40 flanges ....**each $17.50**

Bulkhead Feed-thru Assembly $15.00

Pressure Gauge Section with 15 lb. gauge $10.00

Directional Coupler. UG-40/U Take off 20db $17.50

MAGNET AND STABILIZER CAVITY For 2J41 Magnetron $24.50

90 degree elbows. "E" or "H" Plane 2½" radius $8.50

ADAPTER, waveguide to type "N", UG 81-U, p/o TS 12, TS-13, Etc. $7.50

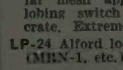
ADAPTER. UG-163/U round cover to special BTL, Flange for TS-45, etc. ....$2.50 ea.

COAXIAL R.F. FILTERS

F-29/SPR-2. Hi-Pass., with 1000 mc. Cut-off. Type "N" input and output. **50 Ohms Z** $9.50

F-41/SPR-I. Hi-Pass, with 300 mc cut-off. Type "N" input and output. **50 Ohms Z** $10.50

MICROWAVE ANTENNAS



3 cm. Horn, 1" x ½", with twist and 180 deg. bend. With dielectric window ....**As shown** $22.50

AT49/APR—Broadband Conical, 300-3300 MC. Type N Feed $8 95

Discone Antenna, AS 125 APR. 1000-3200 mc. Stub supported with type "N" Connector $14.50

ASI4A/AP. 10 CM pick up dipole assy. complete w/ length of coax and "N" connectors $4.50

AS46A/APG-4 Yagi Antenna, 5 element array ..$22.50

30" Parabolic Reflector Spun Aluminum dish $4.85

AN/APA-12—Sector Scan adaptor for APS-2 radar—Complete Kit $37.50

TPS-3, 10 Ft. Dish, "Chicken Wire" Parabola. Extremely lightweight, portable $125.00

AN-154 3 vertical dipoles working against a rectangular mesh approx. 3'x4'. Freq. 140-200 mc. with lobing switch (115v. 60 cy) and portable slatted crate. Extremely rugged $27.95

LP-24 Alford loop, for use with glide-path transmitters (MRN-1, etc.) 100-108 mc. $32.50

THERMISTORS

D-164699 Bead Type DCR, 1525-2550 Ohms @ 75 Deg. F. Coefficient: 2% Per. Deg. Fahr. Max. Current 25 MA AC/DC $1.00

D-167332 Bead Type DCR is 2525-2550 Ohms. Rated 25 MA at 825-1.175 VDC $1.00

D-167613 Disk Type DCR: 355 Ohms @ 75 Deg. F.P.M. 2.5%, 1 Watt $1.00

D-166228 Disk Type 7120 Ohms @ 66° F. 4220 Ohms @ 80°F. 2590 Ohms @ 100°F., 1640 Ohms @ 120°F. $1.00

VACUUM TUBES

IN21	$ .28	5J33	$ 4.75	700D	$ 8.75
IP5GT	.45	5J30	14.50	703A	1.50
IP30	1.10	C5B/5C30	1.10	704A	.95
2C21	.35	C6J/5C21	4.50	705A	.75
2C22/7193	.07	7C4/1203A	.18	706A	9.75
2C26A	.08	9GP7	3.45	706D	14.75
2J21A	2.25	10Y	.10	706EY	9.75
2J22	2.50	15R	.15	706CY	9.75
2J26	2.50	39/44	.08	708A	2.10
2J27	3.00	QK59	39.50	709A	2.45
2J29	18.50	QK60	35.00	713A	.85
2J31	13.95	QK61	50.00	C-722A	.90
2J32	14.50	QK62	54.00	725A	write
2J37	9.50	CEQ72	.59	730A	8.50
2J38	9.00	ML-100	69.50	800	.65
2J39	8.25	HY114B	.25	801	.25
2J48	22.50	227A	2.50	837	.85
2J56	48.50	268A	2.25	843	.19
2J62	6.75	316A	.50	861	12.00
3EPI	1.75	355A	12.50	864	.19
3FP7	1.10	356B	10.50	876	.75
4J34	23.50	393A	4.50	884	1.10
4J38	85.00	417A	6.10	CKI005	.35
4J42	47.50	GL471A	2.10	1625	.20
5FP7	1.10	WL531	2.75	1619	.15
5GP1	4.50	532/1B32	1.10	1626	.10
5HP4	3.50	GL559	.75	1629	.10
5J23	34.50	700B	8.75	8012	1.75

POWER TRANSFORMERS

COMBINATION—115V/60~INPUT

CT-133	150-C-150V/65MA, 6.3V/2.5A, 6.3V/0.6A	$1.79
CT-127	900V/25MA PK. 5V/2A, 2V/7.5A	2.79
CT-006	350-0-350V/120MA, 5VCT/3A, 2.5VCT/12.5A, 2.5VCT/3.5A	4.39
CT-965	78V/0.6A, 6.3V/2A	1.95
CT-004	350-0-350V/90MA, 5VCT/3A, 2.5VCT/12.5A	4.60
CT-002	350-0350V/50MA, 5VCT/2A, 2.5VCT/7.5A	3.65
CT-479	7000V/.018V, 2.5V/5A/17,800 V. Test ..	22.50
CT-013	450-0450V @ 200MA, 10V/1.5A, 2.5, 3.5A, 5V/3A	4.35
CT-403	350VCT .026A 5V/3A	2.75
CT-931	585VCT .086A 5V/3A, 6.3V/6A	4.25
CT-929	4200V/.001A, 2.5V/2A, 6.3VCT/.6A ...	5.35

PLATE—115V/60~INPUT

PT- 07	400VCT/4.0 AMPS For RA43	17.50
PT 034	125V/45MA (For Preamp)	1.15
PT 157	660-0-660VAC (500VDC) or 550-0-550 VAC (400VDC) at 250 MADC	8.70
PT 167	1400-0-1400 VAC (300MADC) or 1175-0-1175 VAC (1000VDC) at 300 MADC	22.50
PT 168	2100-0-2100 VAC (1750VDC) or 1800-0-1800 VAC (1500 VDC) at 300 MADC	33.00
PT 371	210-0-210V at 2.12 Amp.	9.45
PT 133	3140/1570V. 2.36KVA	85.00
PT 801	22,000V/234 MA., 5.35 KVA	115.00
PT 521	7500V/.06A. Half Wave	59.50
PT 913	2500V/12 MA H'SLD	4.95
PT 12A	280VCT/1.2A	3.95
PT-38-2	37.5/40V AT 750 MA	2.15

FILAMENT—115V/60~INPUT

FT-157	4V/16A, 2.5V/2.75A	2.95
FT-101	6V/.25A	.79
FT-924	5.25A/21A, 2x7.75V/6.5A	14.95
FT-824	2x26V/2.5A, 16V/1A, 1.2V/7A, 6.4V/10A 6.4V/2A	8.95
FT-463	6.3VCT/1A, 5VCT/3A, 5VCT/3A	5.49
FT-55-2	7.2V/21.5A, 6.5V/6.85A, 5V/6A, 5V/3A	8.95
FT-38A	6.3V/2.5A, 2x2.5V/7A 5KV Test	2.79
FT-650	2.5V/10A-3KV TEST LO-CAP	7.50
FT-025	2.5VCT/10A, 10KV TEST	6.95

FILTER CHOKES

Stock	Description	Price
CH-914	12HY/250 MA 2500 V. Test	$3.25
CH-CEC 117	9-60H/.05-400 MA. 10 KV Test ...	$14.95
CH-113	2.5H/700 MA, 2.5 KV Test 18 Ohms	5.75
CG-044	8.5H/350 MA, 3.5 KV Test 50 Ohms	6.35
CH-291	0.1H/12 A, DCR: 0.3 Ohms	12.50
CH-322	.35H/350 MA—10 Ohms DCR	2.75
CH-141	Dual 7H/75 MA, 11H/60 MA	4.69
CH-69-1	Dual 120H/17 MA	2.35
CH-8-35	2x.5H/380 MA/25 Ohms	1.79
CH-776	1.28H/130 MA/75 ohms	2.25
CH-344	1.5H/145MA/1200V Test	2.35
CH-43A	10HY/15 MA—850 ohms DCR	1.75
CH-366	20H/300 MA	6.95
CH-999	15HY/15 MA—400 ohms DCR	1.95
CH-445	0.5HY/200 MA, 32.2 ohms, 3000 V.T.	1.39
CH-170	2x0.5H/380 MA, 25 ohms	2.79
CH-533	13.5H, 1.0 AMP DC, 13.5 KVINS ..	29.95

I. F. AMPLIFIER STRIPS

Model 15: 30 Mc center frequency. Bandwidth 2.5 Mc. gain figure: 65 db. Uses 5 stages of 6AC7's. Has D. C. Restorer and Video Detector. A.F.C. Strip included. Input impedance: 50 ohms. Less tubes $17.50

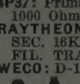
60 MC. Miniature IF strip, using 6AK5's 60 Mc center Freq. Gain: 95 db at Bandwidth of 2.7 Mc. New. Complete with tubes. $15.00

PULSE TRANSFORMERS

4P37: Primary, 50 Ohms 750V, .001 Duty. Sec. 15KV, 1000 Ohms Impedance. Bifilar 12.6V/2.5A $32.50

RAYTHEON WX 4298E: Primary 4KV., 1.0 USEC. SEC. 16KV-16 AMP DUTY RATIO: 001 400 CYCLE FIL. TRANS. "BUILT-IN" $42.50

WECO: D-163247 For Modulator of SCR 720 ....$22.50



GE #K-2449A
Primary: 9.33 KV, 50 ohms Imp.
Secondary: 28 KV, 450 ohms.
Pulse length: 1.05/5 usec @ 635/120 PPS, PK Power Out: 1,740 KW
Bifilar: 1.5 amps (as shown) $62.50

GE #K-2748-A, 0.5 usec @ 2000 Pps. Pk. Pwr. out is 32 KW impedance 40:100 ohm output. Pri. volts 2.8 KV Pk. Sec. volts 11.5 KV Pk. Bifilar rated at 1.3 Amp. Fitted with magnetron well $24.50

K-2745 Primary: 3.1/2.8 KV, 50 ohms Z. Secondary: 14/12.6 KV 1025 ohms Z. Pulse Length: 0.25/1.0 usec @ 600/600 PPS. Pk. Power 200/150 KW. Bifilar: 1.3 Amp. Has "built-in" magnetron well $32.50

K-2461-A. Primary: 3.1/2.6 KV—50 ohms (line). Secondary 14/11.5 KV—1000 ohms Z. Pulse Length: 1 usec @ 600 PPS. Pk. Power Out: 200/130 KW. Bifilar 1.3 Amp. Fitted with magnetron well $29.50

K35J45—Pulse Inversion: PRI: 5 KV PK. Pulse Negative. Sec: Pos. Pulse, 4 KV; 1 usec. and .001 DUTY RATIO $6.50

54J318-1—3 wdgs. Ratio: 1:1:1, 1.10 uh. /wdg., 2.5 ohms DCR $3.50

UTAH X-151T-1: Dual Transformer, 2 Wdgs. per section 1:1 Ratio per sec 13 MH inductance 30 ohms DCR $5.00

UTAH X-150T-1: Two sections, 3 Wdgs. per section, 1:1:1 Ratio, 3 MH, 6 ohms DCR per Wdg. $5.00

68G711: Ratio: 4:1 ri: 200V, Sec. 53V, 1.0 usec Pulse @ 2000 PPS, 0.016 KVA $4.50

TR1049 Ratio 21 Pri. 220 MH, 50 Ohms, sec. 0.75 H. DCR 100 Ohms $6.75

K-904695-501: Ratio 1:1, Pri. Imp. 40 Ohm. Sec. Imp. 40 Ohms. Passes pulse 0.6 usec with 0.05 usec rise $8.95

RAYTHEON, UX-7350, UX-7307 ....$5. ea.

PULSE NETWORKS

7-5E3-1-200-67P, 7.5 KV "E" Circuit, 1 microsec, 200 PPS, 67 ohms impedance 3 sections $7.50

7-5E4-16-60, 67P, 7.5 KV "E" Circuit, 4 sections 16 microsec. 60 PPS, 67 ohms impedance $15.00

7-5E3-3-200-67P, 7.5 KV. "E" Circuit, 3 microsec. 200 PPS, ohms imp. 3 sections $12.50

H-616 10KV, 2.2 usec., 375 PPS, 50 ohms imp..$27.50

H-615 10KV, 0.85 usec., 750 PPS, 50 ohms imp..$27.50

KS8865 CHARGING CHOKE: 115-150 H @ .02A. 32 40H @ .08A, 21 KV Test $37.50

G.E. 25E5-1-350-50 P2T, "E", CKT, 1 Microsec. Pulse @ 350 PPS, 50 OHMS Impedance $69.50

KS9623 CHARGING CHOKE: 16H @ 75 MA, 380 Ohms DCR, 9000 Vac Test $14.95

H-605 25KV, "E" CKT, 1.5 usec. 400 PPS, 50 Ohms Impedance, 5 sections $62.50

PULSE MODULATORS

MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power 144 KW (12 KV at 12 Amp.) Duty Ratio: .001 max. Pulse duration: 5, 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 cps. Uses: 1-71B, 4-89-B, 3-'72's, 1-'73. New ....Less Cover—$135

ASD Modulator-Units, mfd. by Sperry. Hard tube pulser delivers Pk. pulse of 144 kw. Similar to Mod 3 unit. Brand new. less tubes $85.00

Airborne RF head, model A1A, delivers 50 Kw peak output at 9000 mc. at .001 duty. Complete with pulser unit and all tubes $185.00

DELAY NETWORKS

D-168184: 0.5 usec. up to 2000 PPS, 1800 Ohms ..$4.00

D-170499: 0.25/.5/.75usec. 8 KV., 50 Ohms $8.50

D-165997 Delay 1.25 usec ....$ .50

RCA #255686-502: 1.7 usec. 1400 ohm impedance $2.00

D-162311, Delay of 0.5 usec. 72 ohms with 4 MC. Bandwidth $4.75

4-172578, 416 ohms imp., 0.22 usec. Delay $4.75

D-150979: Oscillating network. Oscillates at 81,955 kc. When normal current of 10ma. is interrupted. Has built-in temperature control for stability. Assembled in shielded can 4" L x 4" Diam. $4.35

DYNAMOTORS

	INPUT		OUTPUT		
TYPE	VOLTS	AMPS	VOLTS	AMPS	Price
35X.059	19	3.8	405	.095	$4.35
POSX-15	14	2.8	220	.08	8.95
DA-7A	28	27	1100	.400	15.00
DM 33A	28	7	540	.250	3.95
23350	27	1.75	285	.075	3.95
B-19	12	9.4	275	.110	6.95
			500	.050	
DA-3A*	28	10	300	.260	6.95
			150	.010	
			14.5	5.	
PE 73 CM	28	19	1000	.350	17.50
BD 69‡	14	2.8	220	.08	8.95
DAG-33A	18	3.2	450	.06	2.50
DM 25†	12	2.3	250	.05	6.95
BDAR 93	28	3.25	375	.150	6.95

‡ Less Filter.
† Used, Excellent.
* Replacement for PE 94.

PE 94-, Brand New 5.95

INVERTERS

800-1B Input 24 vdc. 62 A. Output: 115 V. 800 cy. 7A. 1 phase. Used. excellent $18.75

PE-218H: Input: 25/28 vdc. 92 amp. Output: 115V 350/500 cy 1500 Volt-ampere. NEW $32.50

MAIL ORDERS PROMPTLY FILLED. ALL PRICES F.O.B. NEW YORK CITY. SEND M.O. OR CHECK. ONLY SHIPPING SENT C.O.D. RATED CONCERNS SEND P.O. PARCELS IN EXCESS OF 20 POUNDS WILL BE SHIPPED VIA CHEAPEST TRUCK OR RAILEX.

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PROFESSIONAL SERVICES

News-New Products

(Continued from page 194A)

inductance, free from non-linear effects.

Capable of dissipating 2.5 watts with a temperature rise of 20°C, the units have voltage ratings of 400 volts maximum, dimensions of $1\frac{1}{8}$ inches long by $\frac{3}{4}$ inches diameter, weigh approximately one ounce. The same range of inductance values and other corresponding characteristics can be provided with approximately 10 per cent higher Q's when mounted in slightly larger cans. Style A Type 1 Variable Inductors are also available in the form of complete sealed tuned circuits with shunt capacitors included to customer's specifications.